

Philips Technical Review

DEALING WITH TECHNICAL PROBLEMS
RELATING TO THE PRODUCTS, PROCESSES AND INVESTIGATIONS OF
THE PHILIPS INDUSTRIES

PRINTED WIRING IN RADIO SETS

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*The manufacture of most mass-produced articles is highly mechanized, that is to say the conveyance of components and product, the assembly of the components and other operations call for little manpower. This is the situation — to confine ourselves to the radio industry — in the fabrication of thermionic valves *). The assembly of radio sets, on the other hand, was relatively little mechanized until a few years ago, although it had long been done on the conveyor belt and wide use made of electrically driven screwdrivers and similar tools. The mounting of components, the wiring and the soldering of the connections, however, were almost entirely manual operations. This state of affairs has undergone a radical change since the introduction of "printed wiring", which is increasingly coming into use in radio and television sets.*

The assembly of a radio chassis comprises the following operations:

- 1) mounting and fixing the components;
- 2) wiring up the circuit;
- 3) soldering the connection points;
- 4) electrical trimming;
- 5) testing the complete assembly.

The mechanization of the wiring operation had already been thought of before the war. The idea was to apply to a sheet of insulating material a pattern of conducting connections arranged in accordance with the circuit required. One method of doing this was based on the use of insulating sheets covered on one side with copper foil; the wiring pattern was imprinted on the foil with protective paint, after which the uncovered copper was etched away. The result was termed "printed wiring", and this term has been retained to cover other processes subsequently developed.

Advantages and disadvantages of printed wiring

The development of printed wiring processes received a marked impetus during the last world war, particularly in the United States. The reasons were shortage of labour, rising wages and the need for speedy production of all kinds of electronic apparatus. Frequently such equipment was more

complicated than pre-war radio sets and the production runs were relatively small. The latter two circumstances militated against the system of manual wiring, for with every new set put into production mistakes in connections are bound to be made in the beginning, especially when the circuits are intricate. In the long run a stage is normally reached where the number of mistakes is acceptably low, but when the production run is fairly small it is scarcely possible to reach that stage. In this respect printed wiring offers a considerable improvement.

Another advantage of printed wiring is that all points to be soldered lie approximately in the same plane. This makes a form of mechanical soldering possible — *dip-soldering*. We shall return to this later.

From the standpoint of pure radio engineering, printed wiring as against conventional wiring has disadvantages as well as advantages, although they are only of secondary importance. The differences are concerned with 1) the material of the mounting

*) See, for example, G. Alma and F. Prakke, A new series of small radio valves, Philips tech. Rev. **8**, 289-295, 1946. An example relating to TV tubes will be found in: A. H. Edens, A method of sealing the window and cone of television picture-tubes, Philips tech. Rev. **19**, 318-323, 1957/58 (No. 11).

plate (ordinarily metal but insulating material in the case of printed wiring), 2) the material between the connections (air in the one case and solid insulating material in the other) and 3) the configuration of the wiring (in three dimensions and in two dimensions respectively).

Since printed wiring is applied to a sheet of insulating material, the screening effect of a metal chassis is lacking. A favourable point, however, is that the capacitances with respect to earth are smaller. Also there is the advantage of well-defined earth leads (in contrast with the erratic paths through a metal plate) but against this there is the draw-back of their somewhat higher impedance.

The solid insulating material of a printed circuit makes the capacitances between the connections a little higher, but this can be remedied by applying screening strips. Moisture can cause some conduction along the surface, and this must be taken into account when designing the wiring layout.

The two-dimensional configuration of the wiring as opposed to the conventional three-dimensional configuration also has its good and bad aspects, as tabulated below.

<i>Conventional wiring</i> (three dimensions)	<i>Printed wiring</i> (two dimensions)
All critical connections can be short, because they can cross each other.	Some critical connections are always fairly long, having to pass around others.
Capacitances are low mutually and with respect to the chassis, but vary considerably from one set to another	Mutual capacitances are somewhat higher, but very constant from one set to another. Compensations can be introduced in the wiring itself.
Wiring has small surface area but circuit not immediately traceable. Electrical contacts for trimming purposes not always easy to make.	Surface area necessarily about twice as large, but circuit arrangement very distinct and neat. All points readily accessible.

Long connections in printed wiring can sometimes be avoided by using bridging pieces known as jumpers. These are ordinary bent copper wires mounted in the same manner as components, e.g. resistors. Certain components can also often be mounted in such a way that they act as a "bridge".

An important point to consider is the introduction of modifications, as is frequently necessary in the case of a set newly put into production. In sets with orthodox wiring it is usually a simple matter to add a small component or to replace a component

by another of different dimensions. With printed wiring, however, this usually involves considerable difficulties.

Printed wiring was first employed at Philips in 1955 in a small model radiogramophone and shortly afterwards in radio sets of the family "B2X67 U". It then gradually came into use in other types of set, and also in television receivers. In the latter not merely the connections but coils, too, are printed so that here is a case of "printed circuits".

Fabrication of printed wiring panels

Within the scope of this article we cannot deal in detail with the various processes by which printed wiring panels are fabricated (an insulating sheet carrying printed wiring is called a panel). We shall only sketch the broad outlines of the way in which these panels are made for a simple type of radio receiver. We shall then follow the further progress of these panels in the production process.

After the wiring layout has been drawn, exactly as required, it is transferred photochemically to a silk screen. The screen is normally permeable to a certain type of resin, but is made non-permeable by the transferred pattern which corresponds to the conducting parts of the circuit under production. The resin is now applied via the silk screen to both sides of a nickel plate, which has previously been very thinly copper-plated. The wiring pattern is thus transferred to this plate; at the "conducting" parts the copper remains bare, whilst the "non-conducting" parts are covered by the resin. The plate is then once again immersed in a copper-plating bath and the copper — which deposits only on the uncovered parts of the base — is allowed to grow to the required thickness (usually slightly thicker than the coating of resin). This being done, the plate is rinsed and then immersed in an adhesive.

Ten sheets of paper, impregnated with a synthetic resin, are now placed one above the other in a press; on top of these is placed the adhesive-coated plate and this is in turn surmounted by ten more sheets of impregnated paper (*fig. 1*). This assemblage is heated during the pressing operation and each stack of paper is "baked" into a single sheet of paper-reinforced plastic. Owing to the adhesive the copper wiring adheres more strongly to the plastic sheet than does the thin copper layer to the nickel. A parting can thus easily be effected at this latter boundary surface, leaving 1) the nickel plate, which can be used again, and 2) two aggregates, each consisting of a sheet of plastic with the deposited system of wiring on it, surmounted by a thin layer of cop-

per. After etching away the thin layer of copper we have two plastic sheets carrying the printed wiring. The thickness of the wiring is about $0.002''$ ($\frac{1}{50}$ mm), the width and spacing round about $\frac{1}{16}''$ (1 to $1\frac{1}{2}$ mm).

Manufacture of a radio receiver with printed wiring

To illustrate the process of manufacturing radio receivers with printed wiring panels we shall describe the fabrication of type B2X67 U sets, which have been in production for some years now. These sets are of simple design, having either one or two wavebands and no power transformer.

ventional chassis, which allows room for the maximum number of components that may be required. The rest of the sets (beginning with the mixer stage) is standard and is constructed entirely with printed wiring (fig. 2a and b). Because of the larger surface area needed for the printed wiring, the cabinet for this set is somewhat bigger than it would otherwise have been.

As can be seen in fig. 2b, the conductors broaden out at the places where they have to make contact with a component. At each such position there is a hole in the panel. The connecting leads or lugs of the components are pushed through the holes, cut

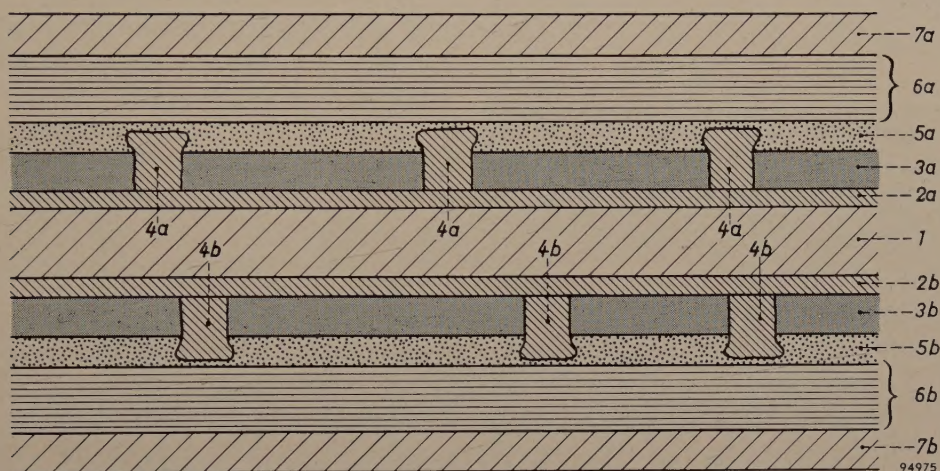


Fig. 1. Schematic cross-section of the stack from which the printed-wiring panels are pressed (not to scale). 1 nickel plate, coated on both sides with a very thin layer of copper 2a-2b. 3a-3b coating of non-permeable resin. 4a-4b electro-deposited copper, which forms the wiring. 5a-5b adhesive. 6a-6b stacks of ten sheets each of impregnated paper. 7a-7b steel platens. After the pressing operation, the two stacks are parted at the boundaries 1-2a and 1-2b, between the nickel and copper.

The simpler model has only one waveband, viz:

Either medium waveband (with "ferroceptor"),
or 16-62 m (with plate aerial),
or 24-94 m (with plate aerial).

It has no tone control.

The more elaborate model has two wavebands:

either medium wave and long wave (with one "ferroceptor"),
or medium wave and 16-62 m (with "ferroceptor" and plate aerial),
or medium wave and 24-94 m (with "ferroceptor" and plate aerial),
or 13-42 m and 41-135 m (with plate aerial only).

This model has a tone control.

Both types can be adapted for a mains voltage (AC or DC), of either 110-127 V or 220 V.

The various models differ considerably in their high-frequency stages; hence each would require a different pattern of printed wiring. This is avoided by mounting the high-frequency stages on a con-

and bent, and are then together dip-soldered to the wiring.

The obvious method of making the holes in the panel is to punch them. However, if an arbitrary pattern of holes of varying sizes is to be punched in the panel, with a different pattern for each type of set, many costly dies would be needed. If, on the other hand, it is possible to use holes of a standard size disposed in a standard grid, a universal stamping tool can be made which can serve for producing any desired combination of holes. There is a tendency nowadays to use holes of $\frac{1}{20}''$ diameter disposed at the intersections of a standard grid of $\frac{1}{10}''$ (2.54 mm) pitch (the holes only being at those intersections where they are needed). A standard grid of this kind imposes certain requirements on the uniformity of the connections of the components used. Various components, such as carbon resistors, paper and

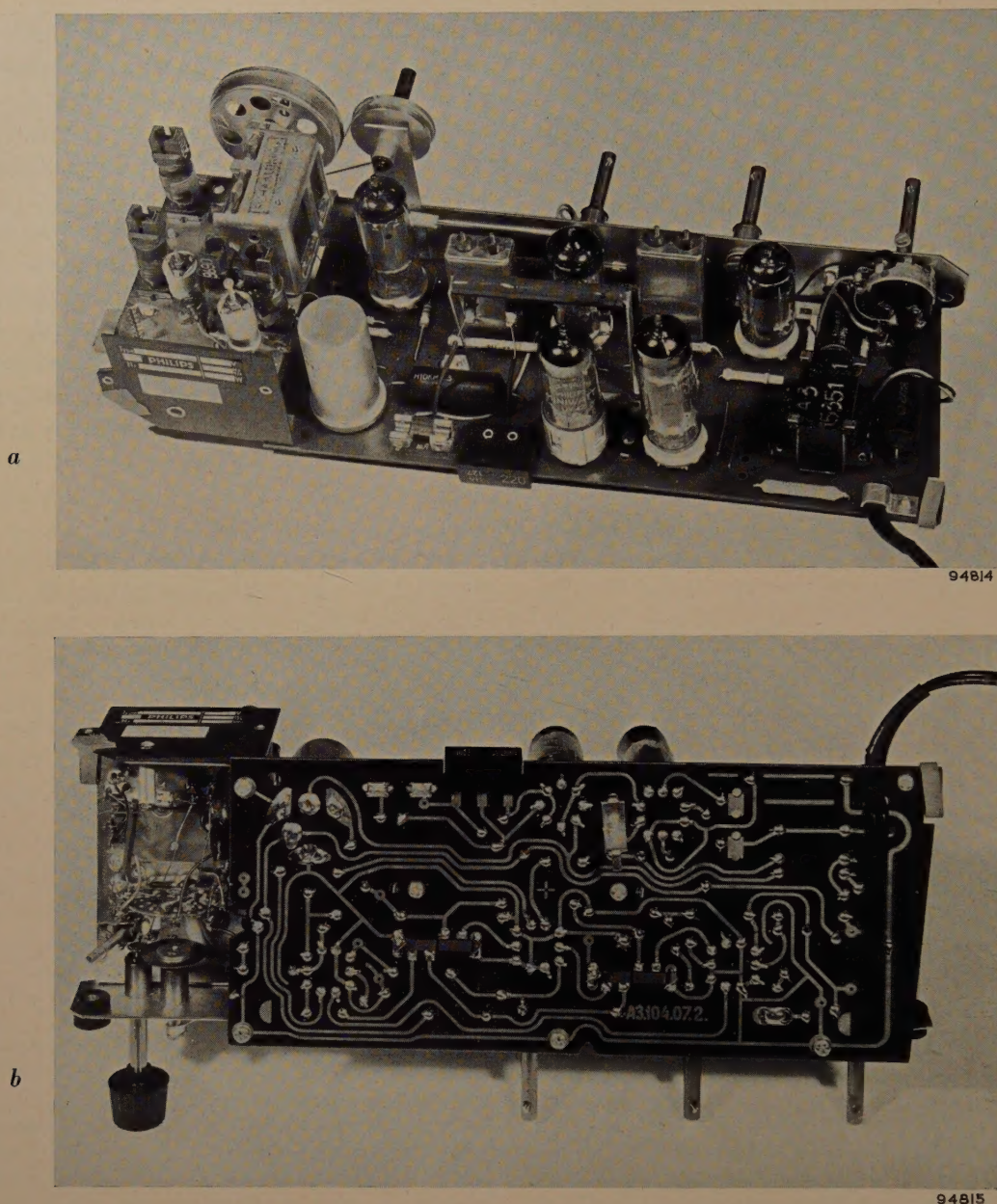


Fig. 2. Chassis of type B2X67 U radio receiver: a) rear top view, b) view from underneath. To the left of the photos is the small conventional chassis carrying the radio-frequency section; the rest of the assembly is the printed wiring panel with its components and valves.

ceramic capacitors, output transformers and some recently developed valve holders, already meet these requirements; others, electrolytic capacitors for instance, do not as yet. The latter, while not precluding the use of printed wiring, call for extra tools both for punching the non-standard holes and for fixing the particular components.

The fabrication of the type of set illustrated in fig. 2 comprises the following operations:

- 1) mounting of components on the panel,
- 2) dip-soldering,
- 3) assembling the high-frequency section (chassis with conventional wiring)
- 4) mounting this chassis to the panel,
- 5) electrical trimming,
- 6) mounting in the cabinet, reception test and packing.

Mounting of components on the panel

After the panels have been thoroughly cleaned, dried and sprayed with soldering flux (fig. 3), the components are mounted in the following order:

- a) components with long connecting leads,
- b) valve holders,
- c) components with "twist" lugs.

a) *Components with connecting leads* include ceramic and paper capacitors and resistors. Jumper wires also come into this category. Before these components are mounted they are subjected to a preparatory operation, which consists in placing 50 or 100 of them in a row on a jig and bending the leads over the jig with a simple tool (fig. 4), such that the centre-to-centre distance of the leads is a multiple of $\frac{1}{10}$ ". The component is then ready to mount on the panel, its two leads being inserted through appropriate holes in the grid pattern. The use of this standard grid limits the number of bending jigs needed.

Racks containing the prepared components are placed within the operative's reach on benches embodying two types of special cutting and bending machines (fig. 5a). The first machine is used only for "thin" components, such as carbon resistors, ceramic capacitors and jumper wires, and the second only for "thick" components, such as paper capacitors. This division into thin and thick components avoids the difficulties that would be involved in exerting simultaneous pressure on both types close together.

The operative places a panel into the machine with the printed wiring facing downwards and inserts the components with their bent connecting leads in the appropriate holes. A pressure pad holds panel and components in position while the machine is operated to cut off the ends of the wires to the required length and bend them over, thereby fixing the components firmly to the panel. The process can be followed with the aid of fig. 6. The panel *M* (fig. 6a) lies on a bronze friction-plate *W*, under which are situated successively a bending plate *B*, a cutter *K* (both of hard steel), and a bronze-clad steel base-plate *S*. The leads inserted through the

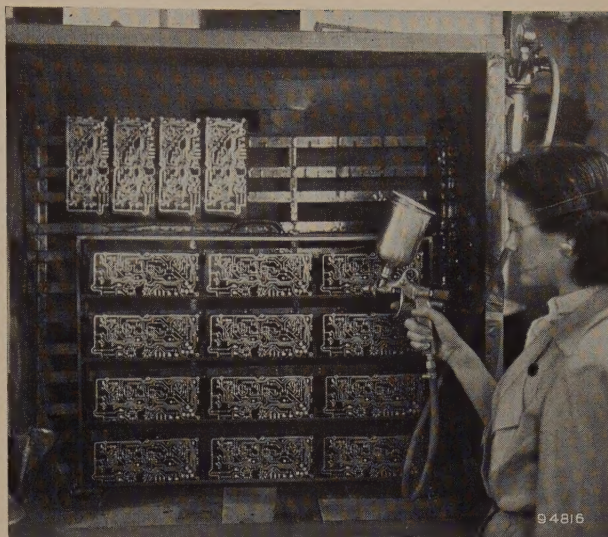


Fig. 3. Spraying soldering flux on the printed wiring.



Fig. 4. Preparation of "thin" components having connecting leads. The leads of 50 or 100 components in a row are simultaneously bent over a jig so that they fit into the holes in the panel.



a

Fig. 5. Cutting and bending machine. a) A panel *M* is placed in the machine and the "thin" components inserted in the panel. b) The cover plate is closed, after which the machine automatically cuts and bends the connecting leads.

b

holes in the panel pass through corresponding holes in plates *W*, *B*, *K* and *S*. Once all thin components are in position, the operative places the pressure pad over them (fig. 5a) and slides forward the cover plate (fig. 5b). The closing of the cover plate initiates the following train of events:

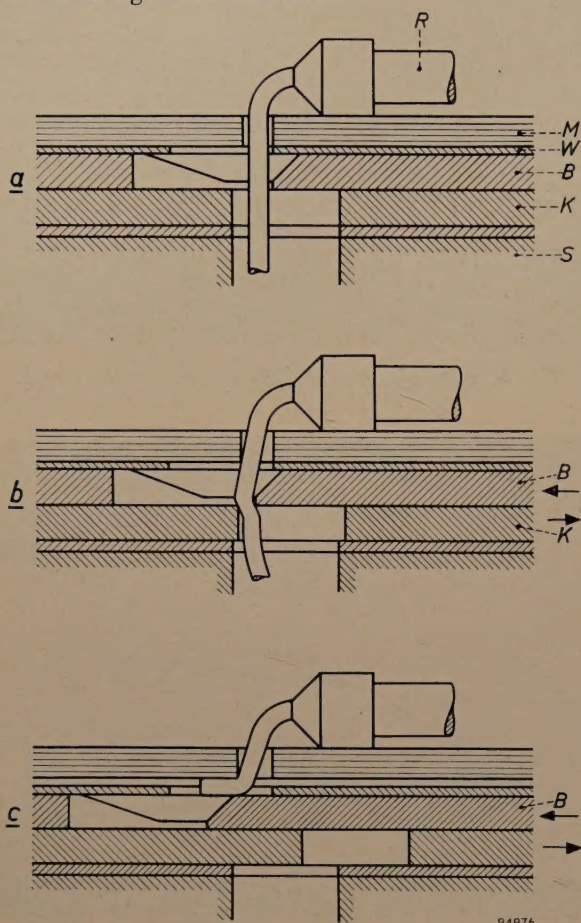


Fig. 6. Cross-section of part of the cutting and bending machine.
a) Starting position. *M* panel with a component *R* (carbon resistor). *W* friction plate. *B* bending plate. *K* cutter. *S* support plate surfaced with bronze.
b) The bending plate moves to the left, the cutter to the right; the left connecting lead of resistor *R* is cut to length.
c) The bending plate has moved farther to the left, bending the lead over against the panel. The right connecting lead is treated in the same way by further movements of the two plates.

- 1) Air under pressure enters the cover plate, the underside of which consists of a rubber diaphragm. The pressure pad is thus pressed down to hold the components firmly in position.
- 2) As soon as the air pressure reaches a certain value, eccentrics set the cutter and the bending plate in motion. As illustrated in fig. 6a-c, this causes the leads to be cut off to the right length and bent over so that the components sit tight.
- 3) The air pressure in the cover plate is released.

The operative now slides the cover plate open, removes the pressure pad and lifts out the panel complete with the thin components.

Among the thick components, whose turn it is to be subjected to the same operations in the second machine, is the output transformer. This has no connecting leads, but short connecting lugs which fit into the grid and are merely bent over in the machine.

The cutting and bending machine is designed to handle a total of 30 components at one time. A larger number of different components could hardly be arranged within reach of the operative. Also the forces involved in the simultaneous cutting and bending of so many leads would become excessive.

The wires connecting the panel to an external component (to the loudspeaker for example) cannot be fixed in the cutting and bending machine, for a straight single wire with one end bent over would fall out of the panel. These wires are bent as shown in fig. 7 underneath the panel by means of a hand-tool, and so prevented from falling out.

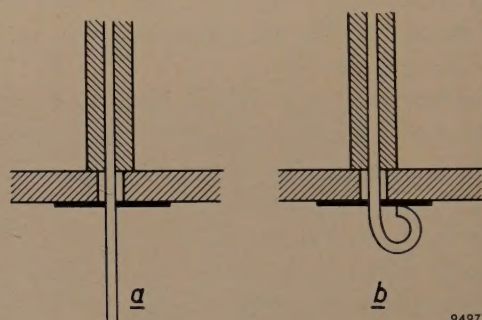


Fig. 7. Method of fixing a wire required to connect a point of the printed wiring to a component outside the chassis (e.g. the loudspeaker).
a) The bared end of the wire is pushed through a hole in the assembly plate.
b) The end is curled over to prevent the wire from falling out. Good electrical contact is effected by dip-soldering.

b) **The valve holders.** A "Noval" base (nine pins at nine corners of a regular decagon, the tenth corner being left open), which is used for many radio valves, does not fit the standard grid of printed wiring panels. For this reason a special valve holder has been designed (fig. 8), the contact sockets of which are provided with lugs (3) arranged to fit in the grid. Protective pads of foam plastic are placed over the lugs when the valve holders are made, to prevent possible damage during transport to the set assembly department.

The valve holders are fixed to the panel with a hand press which simultaneously bends the nine lugs outwards (fig. 8). The press is shown in fig. 9.
c) **Components with "twist" lugs.** The I.F. band-pass filters and electrolytic capacitors have so-called

"twist" lugs, the lugs being pushed through slits in the panel and twisted 60° by a special tool, in the case of the I.F. filters. The capacitor lugs are too short for mechanical twisting and are therefore twisted manually with a pair of pliers.

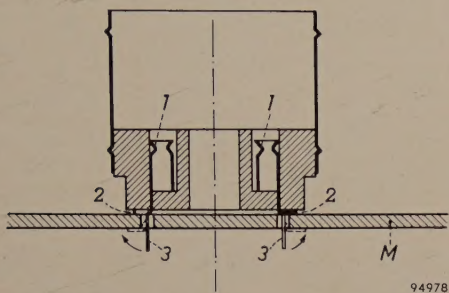


Fig. 8. Holder for valves with "Noval" base. Each contact socket 1 has a split lug. Parts 2 of the lugs are bent round to keep the sockets firmly in the holder. Parts 3 are bent so that they will fit into the standard grid pattern of the panel M. To fix the holder to the panel, the nine lugs 3 are simultaneously bent outwards (see arrows) by a special tool mounted in a hand press.

Dip-soldering

After the panels have once again been sprayed with flux, the electrical connections are soldered. This is done in a bath containing molten solder, the level and temperature of which are automatically kept constant. The panel, printed wiring downwards, is placed on a jig which exposes only those regions where soldering has to take place (fig. 10); the oxide film is brushed from the surface of the molten solder and a safety grill is then closed. The soldering cycle now proceeds mechanically: the jig descends until the underside of the panel just rests on the surface of the solder, and ascends again after a certain time. The panel is kept in vibration both during the dipping process, to ensure good wetting by the solder, and during the ascent, to prevent the formation of unduly large blobs of solder which might establish unwanted connections. Temperature and dipping time are so chosen as to ensure soldered joints of very high quality without allowing the panel to get too hot.

After the connections have been soldered, the valves are fitted in their holders and the assembled panel inspected.

The sequence of operations in the assembly of the panel as described can be seen in the plan in fig. 11. The plan also includes the final stages of the process, which we shall now touch on briefly.

Final stages of assembly

The radio-frequency section of the type B2X67 U receivers consists essentially of a few coils and the variable capacitors, together with trimming capa-

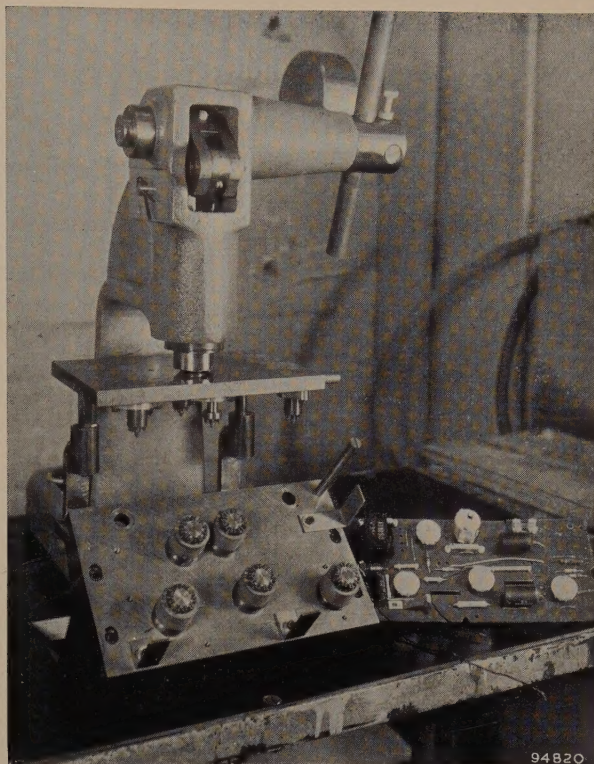


Fig. 9. Hand press for fixing the valve holders to the panels. Foreground, left: lower plate with dies for five valveholders. The upper, punch plate is seen mounted in the press. The dies on the lower plate bend the nine contact lugs outwards (see fig. 8). Right: panel with components, including the five valve holders.

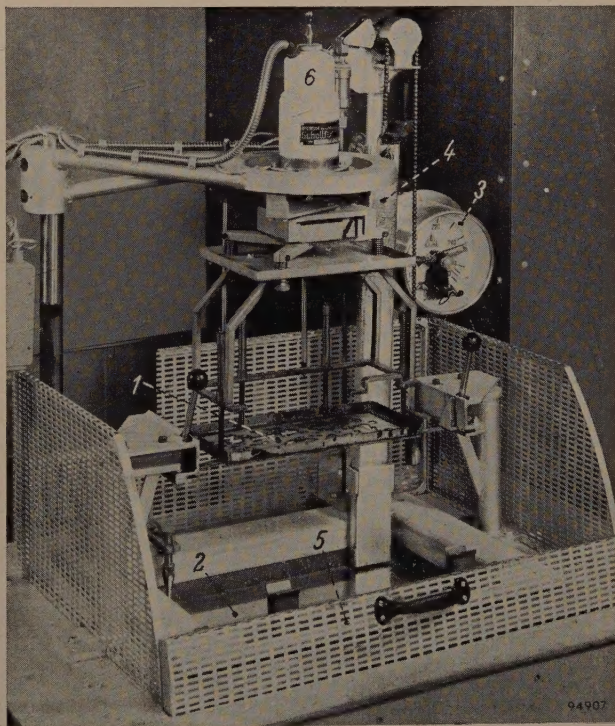


Fig. 10. Dip-soldering machine. The panel, printed wiring downwards, is placed on a jig 1, which shields the parts not to be soldered. 2 bath with molten solder, the temperature and level of which are automatically kept constant. 3 thermometer. 4 bar of solder. 5 safety grill which, when closed, starts the soldering cycle. 6 vibrator, which keeps the panel in vibration during and after soldering.

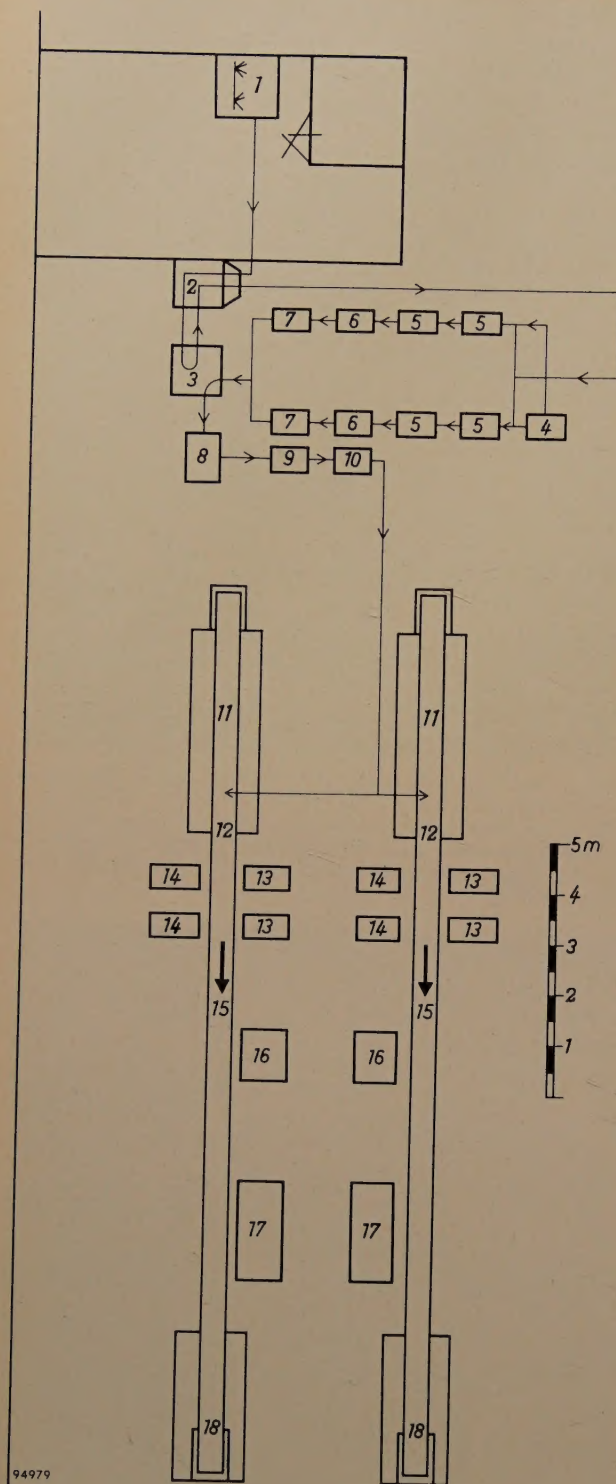


Fig. 11. Plan of assembly department for radio sets incorporating printed wiring. 1 cleaning the printed panels. 2 drying. 3 spraying of soldering flux. 4 preparation of small components. 5 cutting and bending machines. 6 mounting of valve holders. 7 mounting of I.F. band-pass filters and electrolytic capacitors. 8 dip-soldering. 9 insertion of valves. 10 inspection. 11 assembly of R.F. section (metal chassis with conventional wiring) on conveyor belt. 12 mounting of R.F. chassis to printed-wiring panel. 13 trimming. 14 testing and inspection. 15 mounting in cabinets. 16 H.T. test. 17 reception test (R.F. signal modulated with music applied to aerial terminal) in sound-proof cubicle. 18 packing.

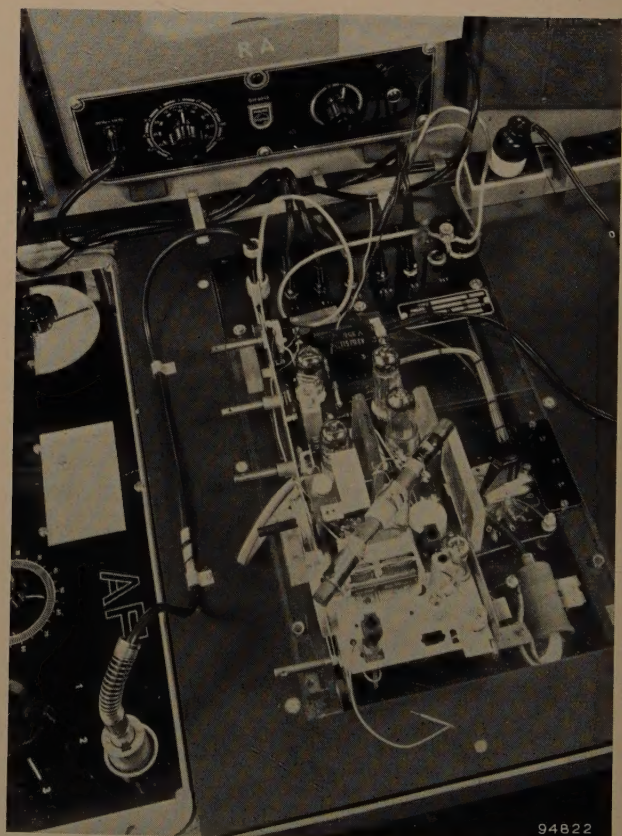
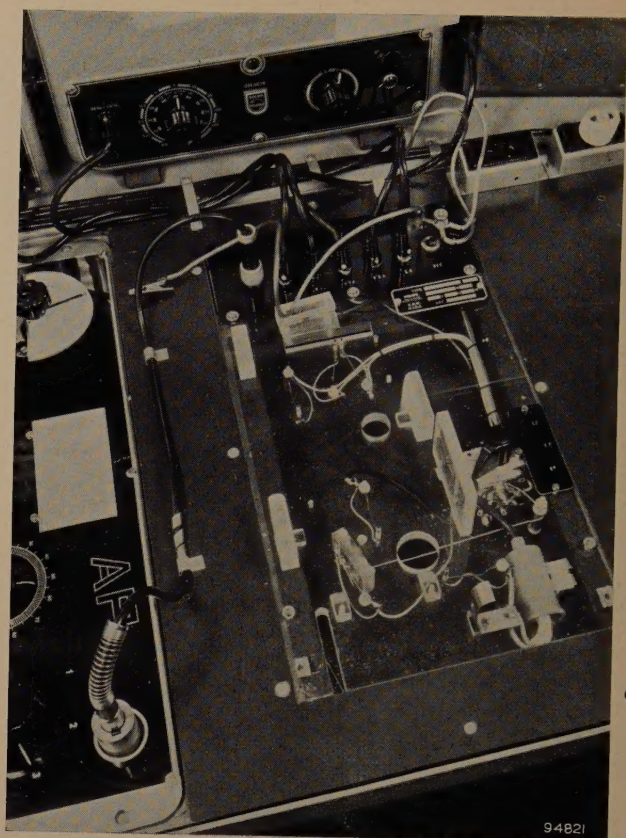


Fig. 12. Test bench for trimming (13 in fig. 11). a) Without chassis. b) With chassis in position. The necessary contacts with the printed wiring are made by spring-loaded pins.

citors and the wave-change switch. These components are mounted and wired up in the conventional way on a metal chassis. As indicated in fig. 11 (11) this is done on a conveyor belt. At the end of the line (at 12), the chassis complete with components is fixed and connected to the completed printed wiring panel to form a single assembly.

In the trimming operations (13) which now follow, advantage is derived from the fact that printed wiring lies in one plane, making every point readily accessible. When the chassis is placed in position on a special test bench, spring-loaded pins make the necessary contacts at various points of the printed wiring for test purposes (fig. 12). Another advantage of printed wiring is that trimming at this stage is required only for the radio-frequency section (the coil and the trimmer capacitor in the aerial circuit and in the oscillator circuit). Because of the reproducible capacitance between the connections on the printed wiring panel the band-pass filters can be adjusted in advance, i.e. during manufacture of these components, and no further adjustments are needed.

The concluding stages — mounting the chassis in the cabinets, the reception test (R.F. signal modulated with music applied to aerial terminal) and packing — are no different from those for receivers constructed by orthodox methods and will therefore not be discussed here. It is worth noting, however, that the factory bay housing the printed wiring section (1...10 in fig. 11) is particularly small compared with the section handling the more conventional assembly work (11...18) and employs relatively few personnel.

Summary. Attempts to mechanize the wiring operations in electronic apparatus have led to the development of "printed wiring". Advantages of this process from the standpoint of organization are that it saves factory space and personnel and practically eliminates the wiring faults which are inevitably made with conventional wiring in the early stages of the production run of a new radio set. From the standpoint of pure radio engineering, printed wiring has both advantages and disadvantages, although these are of only minor significance.

The authors describe briefly the fabrication of printed wiring panels for Philips radio receivers and discuss the manufacture of a simple receiver (type B2X67 U) in which these panels are used.

THE JUNCTION TRANSISTOR AS A NETWORK ELEMENT AT LOW FREQUENCIES

III. STABILIZATION OF THE OPERATING POINT, IN PARTICULAR WITH REGARD TO TEMPERATURE CHANGES

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This, the third and last article on the transistor as a network element at low frequencies, is addressed particularly to readers who are familiar with electronic tubes and who desire to know how the problem of stabilizing the D.C. operating point of a transistor differs from the corresponding problem with valve circuits. The early part of the article is accordingly devoted to explaining how differences between transistors and tubes give rise to differences in circuitry. Here, as in the rest of the article, full use is made of graphical methods. Except in the appendix, formulae relating to the physics of the transistor are not invoked.

The network that (together with the properties of the transistor) determines the D.C. operating point of a transistor calls for much more attention from the designer than is the case for tube circuits. If the properties of the transistor (or the tube) change, that is to say, if the characteristics become changed in shape or are displaced, there is a change in the D.C. operating point. One consequence of this may be that the transistor (or tube) can no longer deliver a signal of the required amplitude; indeed, the output current may be reduced to zero or driven out of the linear range. Fig. 1 illustrates a case that may arise in transistor practice. A D.C. network is assumed to have properties such that 1) I_B , the base current, is maintained at a constant

value of $-50 \mu\text{A}$ which is independent of the transistor properties (this implying that the transistor is biased by a constant current supply to the base) and 2) the operating point Q lies somewhere along the straight line AB . The D.C. operating point is at Q or Q' according as the I_E - V_C characteristic for $I_B = -50 \mu\text{A}$ is represented by the fully-drawn or the broken curve. The maximum amplitude of the A.C. voltage on the output is considerably less in the latter case than it is in the former.

A change of temperature may be regarded as the primary cause of the shift of transistor characteristics. With the OC 71, a shift such as that assumed in fig. 1, whereby the emitter current I_E rises from 2.5 mA to 3.5 mA, may result from a temperature change as small as from 25°C to 40°C . In the second place, the characteristics of two individuals of the same type of transistor may differ considerably. For example, at the same temperature of 25°C , the emitter current for $I_B = -50 \mu\text{A}$ might be 1.75 mA in one OC 71 and 3.75 mA in another individual of the same type.

Generally, it will not be permissible for such large changes to take place in the emitter current. An account of the measures that may be taken to counteract such changes and an analysis of the influence exercised by ambient temperatures will make up the greater part of the article. It is not necessary, for the purpose of the article, to know the physical mechanism underlying the large effect of ambient temperature on the functioning of a transistor, and we shall therefore make no reference to it. It may, however, be pointed out that it is impossible to manufacture transistors that are wholly insusceptible to temperature changes. The underlying cause of temperature effects in semiconductors is the considerable increase with temperature in the equilibrium concentrations of the minority charge-

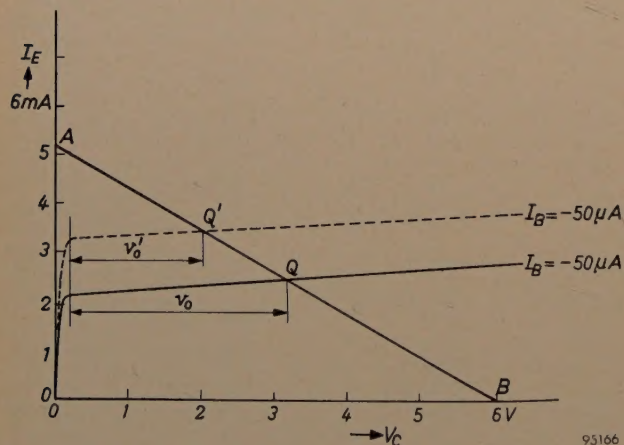


Fig. 1. Displacement of the operating point Q in the output graph of a transistor, in consequence of a change in the transistor characteristics. It is assumed that the characteristic appropriate to a constant base current of $I_B = -50 \mu\text{A}$ moves from the position indicated by the fully-drawn curve to that indicated by the broken-line curve, the shift occurring on account of a rise in temperature, for example, or because the transistor has been exchanged for another individual. If the biasing network responsible for setting the operating point has properties such that the base current maintains the value $-50 \mu\text{A}$ and such that the operating point is forced to remain on line AB , then the point will move from Q to Q' . As a result, the maximum amplitude of the signal voltage across the output will fall from v_o to v_o' .

carriers. Those interested will find a more detailed explanation in the appendix. For the benefit of readers familiar with tube circuits, we shall constantly note points of difference from or similarity to electronic tubes, particularly in the earlier part of the article.

Stabilization of tube and transistor circuits

With tubes the situation is much less complicated than it is with transistors, and for this there are two main reasons. The first is that the ambient temperature has practically no effect on tube characteristics, while it has a considerable effect on those of transistors. The second reason is that, providing there is no flow of grid current, the characteristics of tubes at low frequencies can be displayed by a family of curves drawn in one graph; two graphs are required for displaying transistor characteristics¹⁾. Accordingly, with transistors, we must take account of shifts affecting the characteristics in both graphs.

The operating point P of the tube in the circuit of fig. 2a lies on a vertical straight line $V_g = E_g = \text{constant}$ in the $I_a - V_g$ graph (fig. 3a). The "dynamic" $I_a - V_g$ characteristic, which takes account of the fact that V_a decreases as I_a increases (in consequence of R_a), also appears in this graph. The dynamic characteristic can be derived, in the usual way, from the family of static $I_a - V_g$ characteristics (V_a as running parameter). If for some reason or other — the replacement of the tube, for example, or a change in the anode voltage — this characteris-

tic shifts from position 1 to position 2, the anode current will change by an amount ΔI_a . The commonly employed circuit shown in fig. 2b is a better one, for in this case (see fig. 3a) the operating point lies on a straight line passing through the origin (the tangent of the angle it makes with the ordinate is proportional to R_k), and the change in the anode current is only $\Delta I_a'$. As compared with the circuit

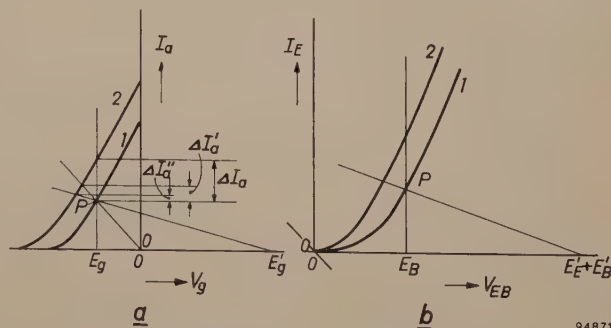


Fig. 3. a) Stabilization of a tube anode current. If curve 1 represents the appropriate $I_a - V_g$ characteristic, then for a certain constant direct voltage E_g on the grid (as in fig. 2a), the operating point will be located at P . If the characteristic shifts into position 2, I_a will change by an amount ΔI_a . With the circuits shown in figs. 2b and 2c, the operating point lies on a straight line passing through the origin or through point E_g' , respectively. Displacement of the characteristic from 1 to 2 results in grid current changes of only $\Delta I_a'$ and $\Delta I_a''$. b) The corresponding graph for a transistor. The characteristics lie on the other side of the ordinate; in consequence of this, the operating point of a transistor in a circuit having a resistor only in the emitter lead (fig. 4b) would be at the origin. Such a circuit is therefore useless.

of fig. 2a, therefore, that of fig. 2b offers a certain stabilization of the anode current against displacement of the $I_a - V_g$ characteristic. The degree of stabilization can be improved by inserting a separate voltage source E_g' in the grid lead (figs. 2c and 3a) with the positive terminal to the grid, and by giving R_k' a correspondingly larger value. This is only done in special cases in tube circuits.

For a meaningful comparison between transistors and tubes the emitter must be regarded as answering to the cathode, the base to the grid, and the collector to the anode. Correlating the electrodes in this fashion is justified by the fact that in a transistor $V_{BC} \approx V_{EC}$ and, if sign be disregarded, $I_E \approx I_C$ ²⁾. Accordingly, we shall make no distinction between V_{BC} and V_{EC} in what follows, but refer only to V_C , calling it the collector voltage. Similarly, we shall refer only to I_E , which we shall simply call the transistor current, without any further qualifica-

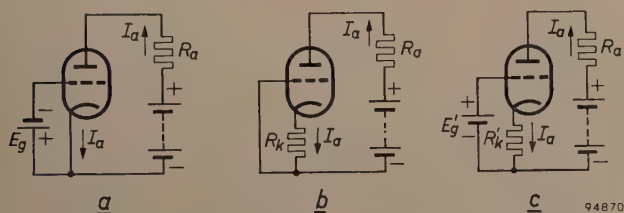


Fig. 2. a) Tube circuit in which the grid is biased with a constant direct voltage E_g . b) Here the grid bias is obtained by means of the voltage drop across cathode resistor R_k . As compared to (a), this circuit provides a certain degree of stabilization of the anode current (see fig. 3a). c) Stabilization of the anode current can be improved by putting a battery E_g' in the grid lead, the polarities being as shown; a larger cathode resistor R_k' can then be used. Alternatively, the battery may be inserted in the cathode lead.

¹⁾ This is discussed in detail in the first of this series of articles, J. P. Beijersbergen, M. Beun and J. te Winkel, The junction transistor as a network element at low frequencies, I. Characteristics and h parameters, Philips tech. Rev. 19, 15-27, 1957/58, particularly pp. 15-16. This article will hereinafter be referred to as I. The graphs discussed in the article (page 18) ought strictly to have a third axis with a scale of temperatures. The curved lines appropriate to given parameter values would then become curved planes.

²⁾ See I, page 16, for a more detailed explanation. At this juncture we would remind the reader of the conventions regarding signs. Currents are denoted as positive when their direction is towards the transistor; as regards voltages, a positive value signifies that the electrode denoted by the first suffix letter has a higher potential than the electrode denoted by the second suffix.

tion. I_E may be taken to stand for either the emitter or the collector current³⁾. The transistor characteristic corresponding to fig. 3a is thus the (dynamic) I_E - V_{EB} characteristic (shown in fig. 3b; it can be derived from the static I_E - V_{EB} characteristics having V_C as their running parameter). The reader will

graph makes it clear that it is a mistake to bias a transistor with a voltage source located between base and emitter (as in fig. 4a): as the operating point now runs on the vertical line $V_{EB} = E_B$, a quite small temperature rise is enough to cause a considerable increase in the transistor current.

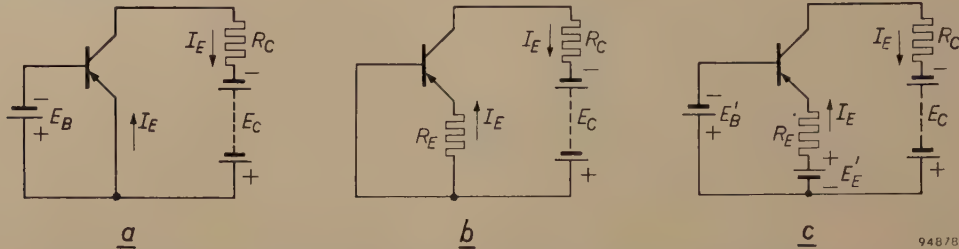


Fig. 4. a), b) and c) are transistor circuits analogous to the tube circuits appearing in figs. 2a, b and c. The circuit (b) with only a resistance R_E in the emitter lead and no voltage source between emitter and base is useless, since the operating-point current is zero (see fig. 3b).

notice straight away that the curve lies on the side of the ordinate opposite to that of the tube characteristics in fig. 3a; this is a further difference between the transistor and the tube that has not so far been mentioned⁴⁾.

The transistor circuits analogous to those of figs. 2a, b and c appear in figs. 4a, b and c. For a circuit as that of fig. 4b, having only a resistor in the emitter lead — and no voltage source between emitter and base — the operating point lies on a straight line passing through the origin. It will be evident from fig. 3b that such a circuit cannot be employed for transistors: I_E would be zero. However, a resistance may well be inserted in the emitter lead provided that a voltage source is inserted between base and emitter (fig. 4c). The voltage source may be placed either in the base or in the emitter lead, and in fig. 4c one has been included in both leads in order to show that there is no preference. It will be seen from the graph in fig. 3b that, compared with fig. 4a, the circuit of fig. 4c offers a certain degree of stabilization of I_E against displacement of the I_E - V_{EB} characteristic.

Fig. 5 shows I_E - V_{EB} characteristics drawn with the transistor temperature t_j (suffix j stands for "junction") as the running parameter. A general rule is that the characteristic moves about 2.2 mV to the left for every °C rise in temperature⁵⁾. The

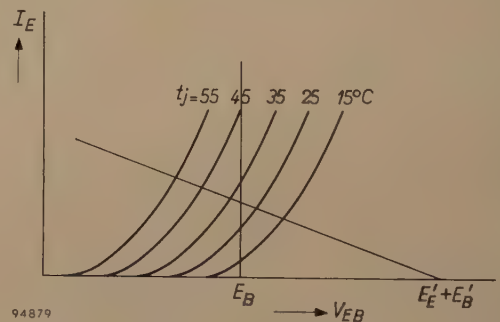


Fig. 5. I_E - V_{EB} transistor characteristics with t_j , the temperature of the transistor crystal, as the running parameter. The characteristic moves about 2.2 mV to the left for every °C rise in t_j . Keeping the base voltage E_B constant (fig. 4a) restricts the operating point to the vertical line, which thus results in a rapid increase of I_E with t_j ; this, then, is an unsatisfactory way of biasing a transistor.

There is a second way of stabilizing I_E against displacement of the I_E - V_{EB} characteristic, this being by leaving out the resistor in the emitter lead (R_E) and putting one in the base lead (R_B). We then have fig. 6. An analogous case does not arise with tubes, for the grid current is zero.

In order to investigate the functioning of this circuit we must consider the relation between I_E and I_B . From now on, therefore, the second graph, that in which I_E is plotted against I_B , will come into the picture (see fig. 7). Applying Kirchhoff's

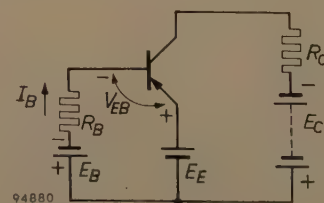


Fig. 6. A second method of rendering the transistor current I_E less sensitive (than it is in fig. 4a) to displacement of the I_E - V_{EB} characteristic is to place a resistor R_B in the base lead instead of R_E in the emitter lead (as in fig. 4c).

³⁾ Without saying so, we have been assuming that the transistor is of the P-N-P type. In this type of transistor the emitter current is positive and the collector current is negative. Choosing the symbol I_E thus does away with unnecessary minus signs. The collector voltage V_C is identified with V_{EC} and V_{BC} (and not with V_{CE} and V_{CB}) for the same reason.

⁴⁾ The difference might also be formulated as follows: in a transistor the potential of the base has a value intermediate between the collector and emitter potentials; in a tube the potential of the grid is not only lower than the anode potential — it is slightly lower even than that of the cathode.

⁵⁾ See appendix.

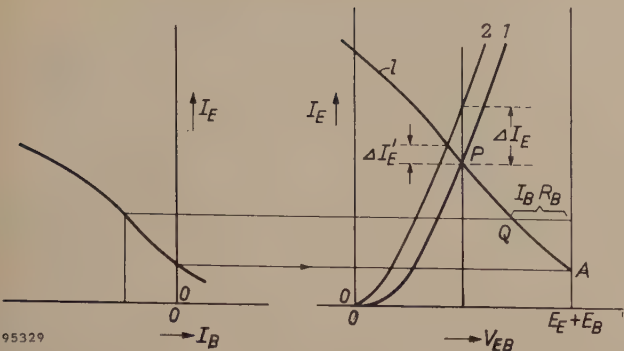


Fig. 7. Location of the operating point P in the I_E - V_{EB} graph for a circuit having a resistor R_B in the base lead (fig. 6). Curve l is constructed by plotting it point by point, use being made of the I_E - I_B graph on the left; P is the point where l intersects the appropriate I_E - V_{EB} characteristic. If this characteristic shifts from position 1 to position 2, V_{EB} being kept constant (fig. 4a), the current increases by an amount ΔI_E ; the effect of resistor R_B is to reduce this increase to $\Delta I_E'$.

second law to the mesh formed by the emitter and base leads in fig. 6, we find that

$$V_{EB} = E_E + E_B + I_B R_B. \quad (1)$$

From the dynamic I_E - I_B characteristic a value of I_B is found that corresponds to some arbitrary value of I_E . (This characteristic too can be derived from the static characteristics, viz. those having V_C as the running parameter.) $I_B R_B$ can then be worked out, whereupon V_{EB} is determined from (1). In this way a point Q in the I_E - V_{EB} graph is obtained (see fig. 7; note that the I_B value associated with point Q is negative). The plotting of a series of such points allows a curve l to be drawn, which represents the locus of the operating point. The curve must, of course, pass through the point A corresponding to $I_B = 0$. It is evident that a high value of R_B and an appropriately large value of $E_E + E_B$ make l more horizontal and therefore provide good stabilization of I_E against displacement of the I_E - V_{EB} characteristic.

We shall now simplify matters by adopting a straight-line approximation to the I_E - I_B characteristic. The equation of the linearized characteristic is

$$I_E = -I_{CE0} - \alpha_{FE} I_B, \quad (2)$$

in which $-I_{CE0}$ is the intercept on the I_E axis (fig. 8) and α_{FE} is the current amplification factor, assumed to be constant, in the common-emitter configuration⁶). Curve l now becomes a straight line through A .

⁶) The symbols employed in this article are those proposed by the Institute of Radio Engineers; see Proc. Inst. Rad. Engrs. 44, 934-937, 1956.
The symbols I_{c0}' and α' are frequently employed in place of I_{CE0} and α_{FE} respectively. Since we are concerned here with dynamic characteristics, it is not entirely correct to use α_{FE} , the static current amplification factor, in equation (2). In doing so we neglect the difference between the static and dynamic characteristics.

The susceptibility of I_E to displacement of the I_E - V_{EB} characteristic diminishes as the value of R_B is increased, but at the same time there is an increase in the susceptibility to displacement of the I_E - I_B characteristic. This can be confirmed by studying fig. 8. We render I_E fully insusceptible to displacement of the I_E - V_{EB} characteristic by employing base current biasing, which implies that I_B is independent of the transistor. The value of I_E associated with this particular value of I_B is fixed by the appropriate I_E - I_B characteristic; thus the I_E - V_{EB} characteristic has ceased to play any part. On the other hand, the susceptibility of I_E to displacement of the I_E - I_B characteristic is now at a maximum. The shift that this characteristic may undergo is considerable. For the OC 71 transistor, an I_{CE0} of $-150 \mu A$ is normal at $25^\circ C$, but it may amount to $-325 \mu A$ in some individuals. The current amplification α_{FE} given by the OC 71 may vary between 35 and 75. I_{CE0} is heavily dependent on temperature too; it is a fairly general rule for germanium transistors that the I_{CE0} value doubles with every $7^\circ C$ temperature rise (see the appendix).

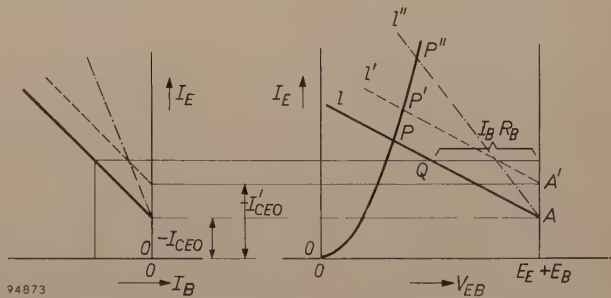


Fig. 8. The same as fig. 7, but here a straight line has been taken as an approximation to the I_E - I_B characteristic. In consequence curve l also becomes a straight line. A is a point on l , and therefore, to be able to draw this straight line, one need only determine one other point Q . If $-I_{CE0}$ (the transistor current at $I_B = 0$) increases, assuming the value $-I'_{CE0}$ say, I_E will also increase; l will shift to l' and P to P' . The operating point also shifts if the slope of the I_E - I_B characteristic alters (e.g. to the position of the chain-dotted line) — in other words, if the current amplification factor alters; line l becomes line l'' and P becomes P'' . The result is again a change in the value of I_E .

The circuit of fig. 9a represents a case of almost pure base current biasing. Here E_B and E_C are zero. The collector voltage is provided by E_E , which therefore has to be large compared to V_{EB} . Hence E_E is approximately equal to $I_B R_B$. Having decided the value I_B is to have, we can thus determine R_B . I_B will now maintain the value chosen for it, irrespective of temperature or the properties of the individual transistor. The circuit is usually drawn in the form shown in fig. 9b. It was this circuit that we had in mind when giving the example

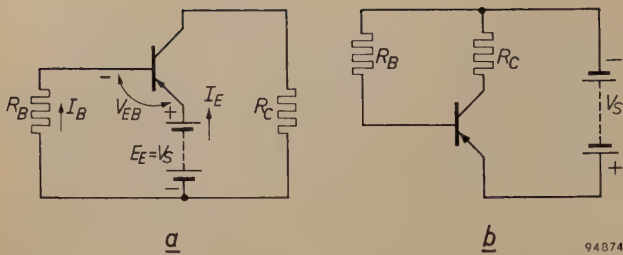


Fig. 9. a) A transistor circuit representing a case of almost pure base current biasing. The transistor current is insensitive to displacement of the I_E - V_{EB} characteristic, but highly sensitive to displacement of the I_E - I_B characteristic. Exactly the opposite applies in the circuit of fig. 4a (which has voltage biasing between emitter and base). b) The same circuit as (a) but drawn in a more conventional form.

of operating-point shift illustrated in fig. 1. Of course, a really pure case of base current biasing only arises if E_E and R_B are made infinitely large, their ratio being kept constant.

Summarizing, one may say that having a voltage source between emitter and base ($R_B = 0$) renders I_E insusceptible to displacement of the I_E - I_B characteristic, but results in maximum susceptibility to displacement of the I_E - V_{EB} characteristic; with a base current biasing ($R_B = \infty$) it is just the reverse. It may be asked which of the two extreme methods is the better. From theoretical considerations (see appendix) it follows that, if the transistor is initially biased such that I_B is zero, the transistor current I_E increases with temperature at about the same rate for both types of biasing. However, as the initial value of I_B is made more and more negative, I_E thereby becoming ever greater, base current biasing becomes more favourable than voltage biasing between emitter and base. It will be seen from fig. 10 that on the whole these inferences hold good in practice. This explains why in discussions of the stabilization of the transistor current it is usually implicit that base current

biasing is employed, this being regarded as the standard case, as it were.

It is also possible to determine the operating-point value of the transistor current in the general case, in a circuit including both R_E and R_B (fig. 11a). Applying once again Kirchhoff's second law to the mesh formed by the emitter and base leads, now in fig. 11a, we find that

$$V_{EB} = E_E + E_B - I_E R_E + I_B R_B. \quad (3)$$

In order to find curve l in the I_E - V_{EB} graph, we assume a certain value for I_E , read the corresponding value of I_B from the I_E - I_B characteristic and proceed to plot $I_B R_B$, starting not from the vertical line $V_{EB} = E_E + E_B$, but from the oblique line

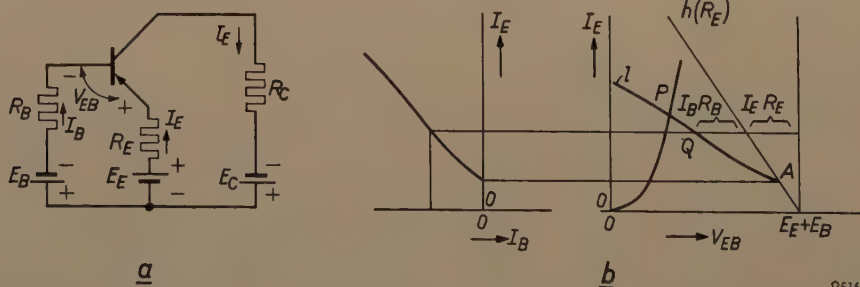
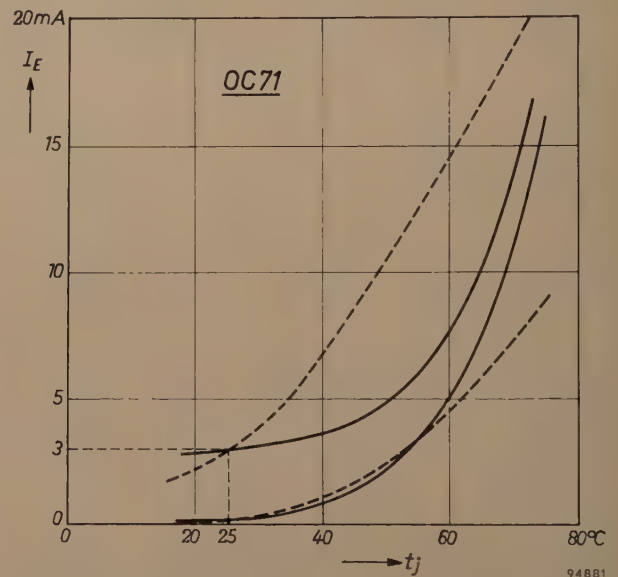


Fig. 11. (a) A transistor circuit with resistors and batteries in the base, emitter and collector leads. (b) Graphical determination of the operating point P for circuit (a). The only difference from the procedure illustrated in fig. 7 is that, in order to find l , one measures the distance $I_B R_B$ not from the vertical line $V_{EB} = E_E + E_B$, but from the line $h(R_E)$, the equation of which is $V_{EB} = E_E + E_B - I_E R_E$. If the I_E - I_B characteristic is taken to be a straight line, curve l will likewise be straight.

$V_{EB} = E_E + E_B - I_E R_E$ (fig. 11b). This new datum line passes through point $E_E + E_B$ on the V_{EB} axis and lies at an angle to the vertical whose tangent is proportional to R_E . (From now on we shall refer to it as $h(R_E)$.) Here too, if the I_E - I_B characteristic is taken to be a straight line, l will also be straight ⁷⁾.

Definition of the D.C. operating point

Of the six variables, viz. three currents (I_E , I_C and I_B) and three voltages (V_{EB} , V_{BC} and V_{CE}) that are associated with a transistor, not more than two can be allotted desired values at a given temperature. The values of the other four quantities are fixed by the two that have been chosen. All this is explained in I on page 15. Similarly, it will generally only be possible to keep two of the said variables constant despite changes in the transistor properties, such as arise on account of temperature changes or because the original transistor has been exchanged for another individual; the values of the other four will usually alter. If therefore it is to be possible to say that two transistors have the same D.C. operating point or that the point remains steady when the temperature changes, one must first decide which two variables are going to be taken to define the operating point. By ignoring the distinction between I_E and $-I_C$ and between V_{EC} and V_{BC} , we have already reduced the number of variables to four, namely to I_E , V_C , I_B and V_{EB} . Of these, we shall select I_E and V_C to characterize the D.C. operating point. There are good reasons for so doing. In the first place, of course, I_E and V_C are the coordinates of the output characteristics for all three of the fundamental configurations, viz. the common-emitter, common-base and common-collector configurations (see I, page 18); the position of the operating point in these graphs determines the range throughout which the transistor can be fully driven (see fig. 1, for example). In the second place, the product $I_E \times V_C$ represents, with fair accuracy, the total amount of power dissipated in the transistor. Hence stabilizing I_E and V_C will ensure that the dissipated power remains within narrow limits, and will remove the danger of the transistor temperature rising too much above ambient temperatures. This matter will be discussed in more detail in the following section. In the third place, the A.C. properties of a given transistor are only slightly dependent on temperature if I_E and V_C are kept constant ⁸⁾.

⁷⁾ This graphical method of determining I_E as a function of t_j was brought to our attention by D. A. Munnik and P. Varekamp, members of the laboratory staff at the Philips Semiconductor Factory at Nijmegen.

Thermal equilibrium

As already stated, the power P_C dissipated in the transistor is approximately equal to $I_E \times V_C$. In a state of thermal equilibrium, the temperature t_j of the transistor crystal will, of course, always exceed the ambient temperature t_{amb} by an amount such that P_C is equal to P_{amb} , the power being transferred from the transistor to its surroundings in the form of heat. A fair approximation for the heat thus transferred away per second is

$$P_{amb} = (t_j - t_{amb})/K, \quad \dots \quad (4)$$

where K is a constant which is the thermal resistance (or reciprocal heat-transfer coefficient), and is mainly dependent on the mechanical design of the transistor and on the manner in which it is mounted on the chassis (i.e., K depends on the thermal contact between the crystal and its surroundings). Data relating to the thermal resistance are supplied by the manufacturer for each type of transistor ⁹⁾. From (4) we see that P_{amb} , plotted as a function of t_j for given values of t_{amb} and K , gives a straight line; see fig. 12.

The properties of the transistor being dependent on t_j , P_C will be some function of t_j which, depending on the circuit and on the transistor itself, may

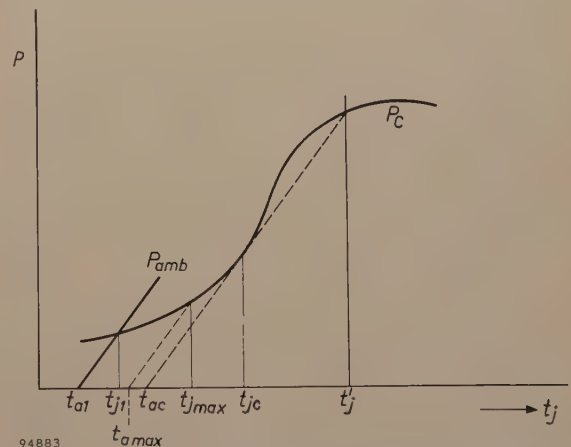


Fig. 12. The power P_C dissipated in a transistor is a function of t_j , the temperature of the transistor crystal. The shape of the function depends on the parameters of the transistor circuit. The power P_{amb} , transferred from the transistor to its surroundings, is proportional to the difference between t_j and the ambient temperature t_{amb} . Provided thermal equilibrium exists, at an ambient temperature of t_{a1} the transistor temperature will be t_{j1} . As t_{amb} increases, the straight line representing P_{amb} moves to the right. If P_C has the shape indicated here, once t_{amb} attains the value t_{ac} the transistor will not remain at t_{jc} but will assume the temperature t_j (temperature "run-away").

⁸⁾ See, for example, J. P. Beijersbergen, M. Beun and J. te Winkel, The junction transistor as a network element at low frequencies, II. Equivalent circuits and dependence of the h parameters on operating point, Philips tech. Rev. **19**, 98-105, 1957/58, particularly Table III and formulae (4) and (5). This article will hereinafter be referred to as II.

⁹⁾ If the transistor is fitted with a cooling fin, its thermal resistance depends to some extent on the ambient temperature.

for example have the form arbitrarily assumed in fig. 12. It is clear that the transistor will only be in thermal equilibrium when its temperature has attained a value t_{j1} at which $P_C = P_{amb}$. The state of equilibrium will be stable, for if t_j rises higher than t_{j1} , P_{amb} will exceed P_C , with the result that t_j decreases; if on the other hand t_j falls below t_{j1} , P_{amb} will be less than P_C , with the result that t_j increases until it attains the value t_{j1} .

As the ambient temperature rises, the straight line P_{amb} is displaced to the right, maintaining the same slope. If in this way it becomes tangential to the curve at a point of increasing slope (broken line, when $t_{amb} = t_{ac}$), the transistor temperature will rise considerably, reaching the value t_j' (temperature "run-away").

The curve of P_C cannot go on rising indefinitely. This is in any case evident from the fact that the D.C. source is unable to supply unlimited power. Usually P_C exhibits a maximum, as has been assumed in fig. 12. It may easily happen, however, that the transistor breaks down before this maximum is attained.

The manufacturer generally indicates a maximum permissible value of transistor temperature (t_{jmax}) because it is impossible to give any guarantee regarding the life of the transistor if it is operated above this temperature. In the assumed case illustrated in fig. 12, the circuit in question can only be employed at ambient temperatures below t_{amax} . In other cases the circuit might be such as to give the P_C curve a quite different shape, in which t_{jc} occurred at a value lower than t_{jmax} ; the ambient temperature would then have to remain below t_{ac} (the intercept of the tangent).

At any given value of t_j , $P_C - P_{amb}$ represents the amount of heat that becomes available each second to raise the temperature of the transistor crystal. Of course, if $P_C - P_{amb}$ is negative, it represents the amount of heat withdrawn from the crystal each second, in which case t_j drops. A state of equilibrium will be attained all the quicker the more rapidly the curves P_C and P_{amb} approach each other. It is not impossible for the two curves to be almost coincident over a considerable range of temperatures in the region of their (stable) point of intersection. If this is so, it will take a long time for equilibrium to be attained; moreover, a slight change in t_{amb} will cause the transistor to start creeping towards a new and distant equilibrium temperature. The new equilibrium temperature may well be higher than t_{jmax} .

We shall now describe how the curve of P_C for a given circuit can be determined, given the transistor characteristics.

Determination of I_E , V_C and P_C as functions of t_j

The D.C. circuits discussed in this article can all be reduced to the fundamental D.C. circuit that appears in fig. 11a. One of the three e.m.f.'s shown

in the diagram can, of course be left out without diminishing the generality. The reason why the three are shown is to make it convenient to compare an actual practical circuit with this fundamental one.

If the $I_E - I_B$ and $I_E - V_{EB}$ characteristics for a given value of t_j are available, the corresponding I_E can be determined by the graphical method illustrated in fig. 11b. If families of these characteristics, corresponding to a set of t_j values, are available, it is possible to determine I_E as a function of t_j . The following formula (Kirchhoff's second law, applied to the mesh formed by the emitter and collector leads in fig. 11a) can then be used for calculating V_C for each I_E value:

$$V_C = E_E + E_C - (R_E + R_C)I_E. \quad (5)$$

We can now plot V_C as a function of t_j ; P_C can also be so plotted, since $P_C = I_E \times V_C$. If K is also known, P_{amb} can also be plotted (eq. 4). We thus have a graph similar to fig. 12 from which we can find out how the circuit will behave as the ambient temperature changes.

As observed earlier in this article, it is the *dynamic* characteristics that should really be used; these take account of the fact that I_E depends on V_C . However, satisfactory results are generally obtained by the use of the static characteristics, valid for constant V_C .

This amounts to making use of straight lines parallel to the V_C axis as approximations to the $I_E - V_C$ characteristics having I_B as the running parameter, as also to those having V_{EB} as the running parameter. The main implication of these approximations is that the *Early effect* is being neglected (see II, page 104).

The required characteristics — those drawn in fig. 11b — can be determined by measurement, the transistor being placed in a thermostatted enclosure. It is advisable to have the collector voltage low (e.g. 2 V) for the purpose of these measurements, in order to keep down the power dissipated in the transistor, so that the transistor crystal remains at the temperature of the thermostat.

If the necessary equipment is not available, one may proceed as follows. The manufacturer supplies $I_E - I_B$ and $I_E - V_{EB}$ characteristics for one temperature (usually $t_j = 25^\circ\text{C}$); alternatively the curves can be worked out from the data supplied. The corresponding curves for other t_j values can be obtained with the aid of two rules that have already been given in the course of this article:

- 1) The $I_E - V_{EB}$ curve moves about 2.2 mV to the left for every $^\circ\text{C}$ rise in t_j .
- 2) The $I_E - I_B$ curve moves upwards by about 10% for every $^\circ\text{C}$ rise in t_j (I_E doubling with every 7°C).

An explanation of these rules will be found in the appendix.

Simplified method

The method just described is proper to an exact analysis of the thermal behaviour of a given transistor in a given circuit, but it is rather involved.

It has already been observed that I becomes a straight line if it be assumed that α_{FE} is independent

of I_E . If it be further assumed that α_{FE} is independent of temperature, the I_E - I_B characteristics having t_j as the running parameter become parallel straight lines, and so do the l curves. There is then no need to draw the I_E - I_B characteristics. All that is necessary is to plot a value $-I_{CE0}(t_{j1})$ on the I_E axis, whence we find the corresponding point $A(t_{j1})$ on the $h(R_E)$ line (fig. 13). A second point

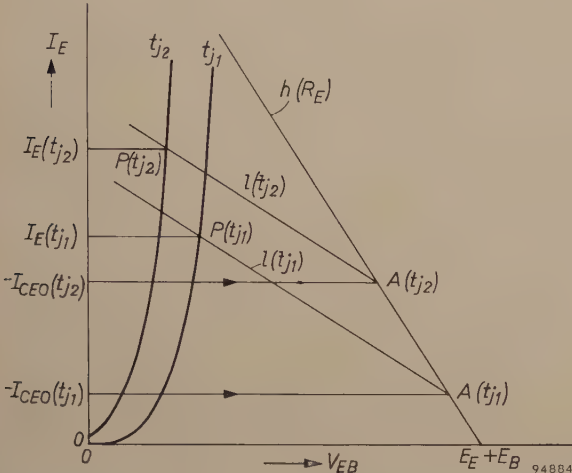


Fig. 13. If α_{FE} , the current amplification factor, is assumed to be independent of I_E and t_j , the l curves become parallel straight lines. It is an easy matter to find the points A at which they intersect with straight line $h(R_E)$, and hence to determine the operating point P for various values of t_j .

can be determined with the aid of α_{FE} (the value of which, of course, has to be known) and equation (2). Taking another value of t_j , say t_{j2} , one proceeds in the same manner to find $A(t_{j2})$, subsequently drawing a line $l(t_{j2})$ parallel to $l(t_{j1})$ through point $A(t_{j2})$.

This procedure is quite satisfactory in cases where I_E does not exceed a few milliamperes. Where currents are higher, the fact that α_{FE} decreases must be taken into account. The current amplification factor of an average OC 72, for example, falls off from 70 to 50 as I_E increases from 10 mA to 80 mA. Usually this kind of information is included in the manufacturer's data. The meaning of α_{FE} in such data is $\alpha_{FE} = \{I_E - (-I_{CE0})\} / I_B$. Often the symbol α' is used to denote this quantity.

Design of a circuit to meet a specified thermal stability

The circuit in fig. 14a is one that is very commonly employed. If the supply (of e.m.f. V_S) has a negligible internal resistance, it will be found by applying Thévenin's theorem¹⁰⁾ that fig. 14a, as seen by the transistor, is equivalent to fig. 14c. This in its turn is equivalent to fig. 11a provided E_E in that figure is taken to be zero and R_B and E_B have the following relationship to the resistances in the circuit of fig. 14b:

$$R_B = \frac{R_1 R_2}{R_1 + R_2}, \quad \dots \dots (6)$$

$$E_B = V_S \frac{R_1}{R_1 + R_2}. \quad \dots \dots (7)$$

We shall now discuss the case of a circuit that is required to have a certain specified thermal stability: we shall suppose that it has already been decided what the D.C. operating point shall be at a certain ambient temperature (V_{C1} , I_{E1} at t_{amb1}), and that it is not permissible for the current to rise higher than I_{E2} at a certain higher ambient temperature t_{amb2} . If V_S , the voltage of the supply source, is given, then

$$R_E + R_C = (V_S - V_{C1}) / I_{E1}. \quad \dots (8)$$

We shall assume that R_C is known, and hence we shall be able to work out R_E . The powers dissipated in the transistor at the two ambient temperatures of t_{amb1} and t_{amb2} are $P_{C1} = I_{E1} \times V_{C1}$ and $P_{C2} = I_{E2} \times V_{C2}$, respectively. V_{C2} , the operating-point voltage at t_{amb2} , is worked out from (8), the suffixes therein being changed from 1 to 2. For a state of thermal equilibrium to exist, P_C must be equal to P_{amb} ; t_{j1} and t_{j2} , the temperatures the transistor

¹⁰⁾ Thévenin's theorem implies that it is permissible to replace the part of the network to the left of the broken line (fig. 14b) by an e.m.f. in series with a resistance. The e.m.f. is equal to the open-circuit voltage across the terminals located on the broken line; the resistance is that measured between these terminals when V_S is short-circuited.

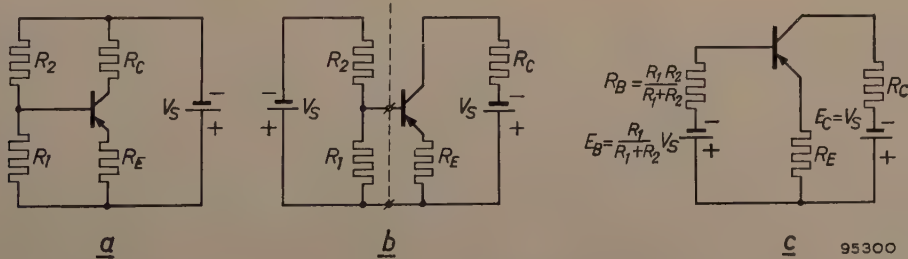


Fig. 14. a) A direct-current circuit frequently employed in transistor practice. If the internal resistance of the battery (of e.m.f. = V_S) is negligible, then, seen from the transistor, (a) is equivalent to (b). Application of Thévenin's theorem¹⁰⁾ to the portion on the left of the dotted line results in the transformation of (b) into (c), the circuit thereby being reduced to that in fig. 11a.

will have when the temperatures of its surroundings are t_{amb1} and t_{amb2} , can therefore be worked out from (4). V_{EB1} and I_{B1} , the values associated with the combination of t_{j1} with I_{E1} , are taken from the curves appropriate to the transistor temperature t_{j1} . Insertion of these values and that of R_E in (3) gives us an equation involving two unknowns, $E_E + E_B$ and R_B . A second equation is obtained by proceeding in the same manner but by taking the combination of t_{j2} with I_{E2} . From the two equations R_B and E_B can be calculated ($E_E = 0$). Finally, in (6) and (7) we have two further equations that allow us to work out the values of resistors R_1 and R_2 . This completes the dimensioning of the circuit.

The foregoing implies the use of the static I_E - I_B and I_E - V_{EB} characteristics. Sometimes a more exact procedure, based on the dynamic characteristics, is desirable. The data supplied by the manufacturer are generally sufficient to allow the static characteristic (with V_C as the parameter) to be derived for one temperature (usually $t_j = 25^\circ\text{C}$). Once $R_E + R_C$ has been worked out from (8), it is possible to construct the dynamic characteristics for this same temperature of $t_j = 25^\circ\text{C}$. The two rules given earlier on page 128 can equally well be applied to these dynamic characteristics, in order to get some idea of their position at other temperatures.

Spread in properties. An example of stabilized design

It was stated in the earlier part of the article that different individuals of the same transistor type may exhibit a considerable divergence in their properties. If a transistor-equipped device is to be produced in quantity, and if it is desired to avoid the need to bias each transistor individually, then allowance must be made for this spread in properties. We shall explain how this is done by working through a simple example; the design will also allow for changes in ambient temperature.

Our example concerns an OC 71 in the driver stage of an amplifier. The circuit of which the OC 71 forms part is as shown in fig. 15a. From the A.C. point of view, the transistor is in the common-emitter configuration (fig. 15b). The performance

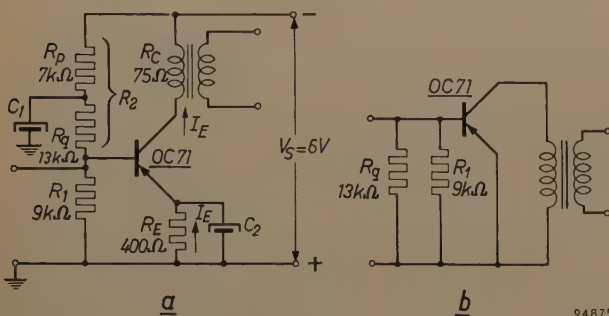


Fig. 15. a) Circuit for a type OC 71 transistor in the driver stage of an amplifier. By way of illustration, the values of the various resistors are worked out in the text in conformity with an assumed set of requirements.
b) A.C. circuit corresponding to (a).

required of the output stage necessitates that the OC 71 of the driver stage should deliver a signal voltage of 3.8 V peak, and a signal current of 2.7 mA peak, at its output. Assuming a knee voltage of $V_C = 0.3$ V, we see that the operating point in the I_E - V_C graph (fig. 16) will have to remain on

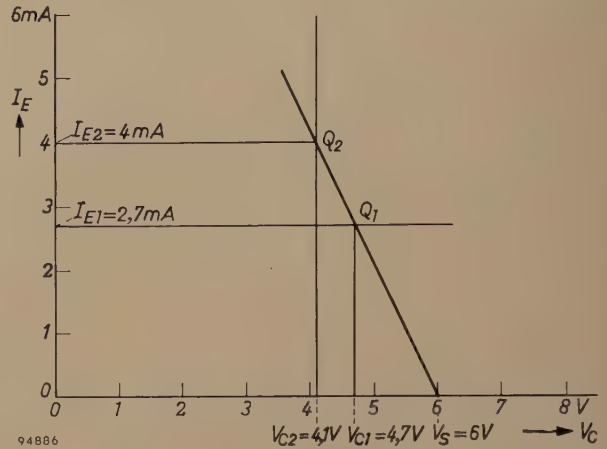


Fig. 16. With regard to the operating point of the transistor in the circuit of fig. 15, it is required that I_E shall remain between 2.7 mA and 4 mA and that V_C shall not drop below 4.1 V. The supply voltage $V_S = 6$ V is given, thus fixing the direct-current load line Q_1Q_2 .

the right of the vertical line $V_C = 3.8 + 0.3 = 4.1$ V and above the horizontal line $I_E = 2.7$ mA. The D.C. resistance of the primary winding of the transformer coupling the driver stage to the output stage is $R_C = 75 \Omega$. The supply voltage is $V_S = 6$ V. It is desired that the transistor current I_E shall not rise higher than 4 mA, since the D.C. magnetization of the transformer would then be excessive. This limitation fixes the D.C. load line in fig. 16. From its slope we find that $R_E + R_C = 475 \Omega$. Since $R_C = 75 \Omega$, we have $R_E = 400 \Omega$. The voltage and current values appropriate to operating point Q_1 are $V_{C1} = 4.7$ V and $I_{E1} = 2.7$ mA. The power then dissipated is $P_{C1} = 4.7 \times 2.7 = 12.7$ mW. For an OC 71 mounted so as to allow the free access of air, K has a value of $0.4^\circ\text{C}/\text{mW}$. When the transistor is operating at Q_1 , therefore, the crystal temperature t_j is $12.7 \times 0.4 \approx 5^\circ\text{C}$ higher than the ambient temperature. Proceeding in a similar way, we find that at Q_2 the crystal temperature is nearly 7°C higher than the ambient temperature. It is required that the amplifier may be used at ambient temperatures between 15°C and 45°C ; in other words, the operating point must keep within Q_1 to Q_2 at these ambient temperatures. Now, a low value of I_E is favoured by a low temperature and low values of α_{FE} and $-I_{CE0}$; on the other hand, a high temperature and high values of α_{FE} and $-I_{CE0}$ tend to result in high I_E values. (It will be seen

in a moment that this is so.) In *fig. 17* I_E - V_{EB} characteristics have been drawn that are appropriate to the lowest and highest values of t_j , namely to $t_{j1} = 15 \pm 5 = 20^\circ\text{C}$ and $t_{j2} = 45 \pm 7 = 52^\circ\text{C}$. I_E - I_B characteristics for these t_j values appear on the left-hand side of the figure. The $t_{j1} = 20^\circ\text{C}$ characteristic is that for minimum values of α_{FE} and $-I_{CE0}$ (the latter being taken to be zero);

and low I_{CE0} and α_{FE} values combine to keep I_{E} low, whereas a high temperature and high I_{CE0} and α_{FE} values have the opposite effects. That this is so can easily be confirmed from fig. 17: as α_{FE} increases, the $l(t_j)$ lines turn clockwise about their points of intersection with $h(R_{\text{E}})$ — these being the points marked A in fig. 13 — and I_{E} increases. (But this is not entirely obvious, as may be seen

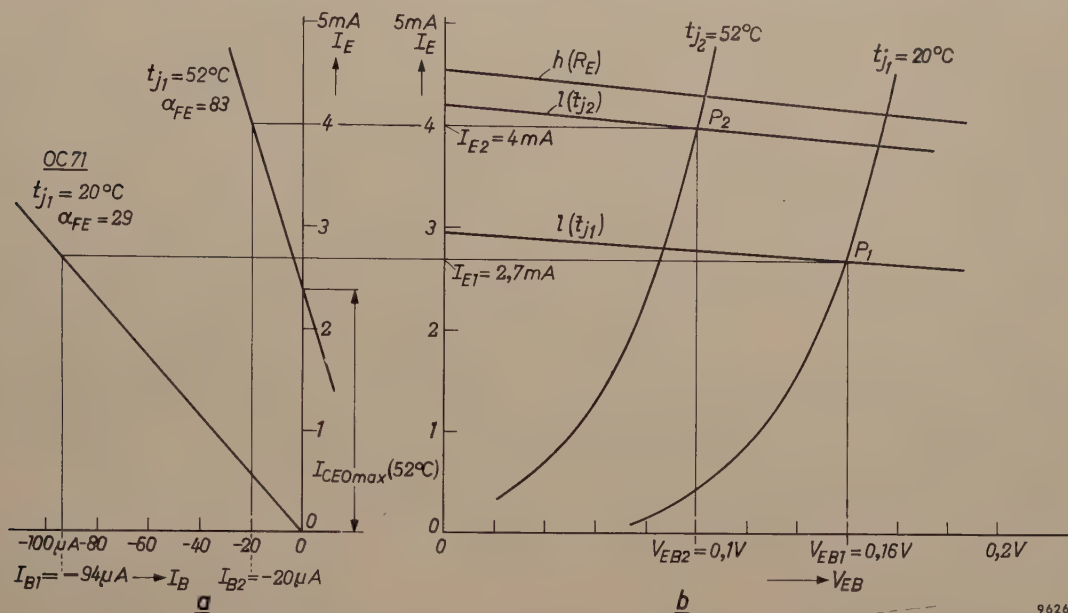


Fig. 17. I_E - V_{EB} characteristics (on the right) and I_E - I_B characteristics (on the left) for an OC 71 at transistor temperatures of $t_{j1} = 20^\circ\text{C}$ and $t_{j2} = 52^\circ\text{C}$. The I_E - I_B characteristic for 20°C is appropriate to an individual transistor whose I_{CE0} and α_{FE} are the lowest occurring in the spread of values; that for 52°C applies to an individual whose I_{CE0} and α_{FE} are the highest occurring. This graph is used to work out the circuit of fig. 15.

the $t_{j_2} = 52^\circ\text{C}$ characteristic is that for maximum values of α_{FE} and $-I_{\text{CE0}}$. Straight lines have been drawn as approximations to the true $I_{\text{E}}-I_{\text{B}}$ characteristics; they are not parallel, however, on account of the difference in the α_{FE} values. Fig. 17 gives us the following information: $V_{\text{EB1}} = 0.16\text{ V}$; $V_{\text{EB2}} = 0.1\text{ V}$; $I_{\text{B1}} = -94\text{ }\mu\text{A}$; $I_{\text{B2}} = -20\text{ }\mu\text{A}$. Determining R_{B} and E_{B} with the aid of (3) in the manner already described (page 130) we find that $R_{\text{B}} = 6.2\text{ k}\Omega$ and $E_{\text{B}} = 1.83\text{ V}$. We now determine R_1 and R_2 with the aid of (6) and (7), and find that $R_1 \approx 9\text{ k}\Omega$ and $R_2 \approx 20\text{ k}\Omega$.

The lines $h(R_E)$, $l(t_{j1})$ and $l(t_{j2})$ are also drawn in fig. 17. They are not necessary for the calculation, but give a general impression of the situation. In the first place they show that, with the scales used in the figure, the point E_B on the abscissa and the points A_1 and A_2 , where $h(R_E)$ intersects $l(t_{j1})$ and $l(t_{j2})$ respectively, lie well outside the graph. It was for this reason that formula (3) was used for determining E_B and R_B and that no attempt was made to arrive at them by geometrical construction.

Secondly, it was stated that a low temperature

from *fig. 18*. Here A is on the left of the appropriate I_E - V_{EB} characteristic and, as a result, I_E decreases instead of increasing as α_{FE} increases.)

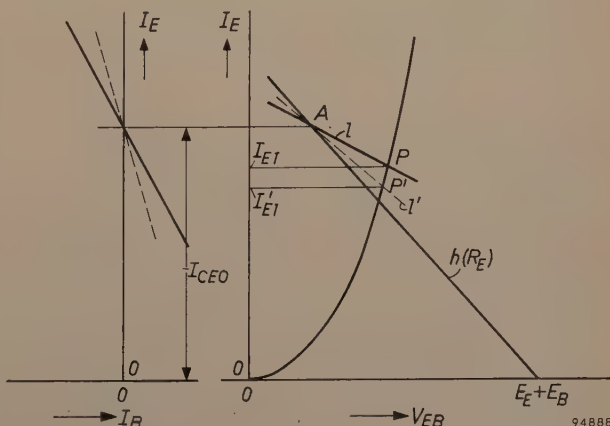


Fig. 18. Demonstration of the fact that it is possible for an increase in a_{FE} to be associated with a fall-off in I_E : this occurs when point A lies on the left of the appropriate I_E-V_{EB} characteristic. As a_{FE} increases, the I_E-I_B characteristic rotates, moving clockwise from a position such as that of the fully-drawn line to one such as that shown by the dotted line. Line l turns about A in the same sense, moving into position l' . The current decreases from I_{E1} to I_{E1}' .

In the third place, fig. 17 reveals that shifting the I_E - V_{EB} characteristic to the right also tends to keep I_E down. This suggests that, to allow for the spread in this characteristic, we ought to have used the curve furthest to the right for $t_j = 20^\circ\text{C}$ and that furthest to the left for $t_j = 52^\circ\text{C}$. However, another glance at the figure will make it clear that the position of the I_E - V_{EB} characteristic has relatively little effect on the value of I_E . We may therefore conclude that our use here of the I_E - V_{EB} characteristics proper to an average transistor will not give rise to any grave error. There are, however, cases in which the position of the I_E - V_{EB} characteristics exercises a considerable influence — particularly where the transistor is a high-power type and heavier currents occur, in consequence of which the calculations arrive at much lower values for the resistors R_E and R_B .

Resistor R_2 (see fig. 15a) is split into resistors R_p and R_q , and the point between them is earthed for alternating current via capacitor C_1 . The object is to prevent the alternating voltages that arise across the internal resistance of the supply source (a quantity that has so far been disregarded) from reaching the input side of the transistor. It will be seen from fig. 15b that, from the alternating current point of view, R_1 and R_q are in parallel across the input of the transistor. The value of R_q might be 13 k Ω , say; the resistance of R_1 and R_q combined would then be 5.7 k Ω . The OC 71 has an input resistance of approximately 1 k Ω . In such a case, where no large currents are involved it is preferable to have the OC 71 driven by a current¹¹⁾ rather than a voltage signal, in order to keep down distortion; it is therefore desirable that the resistance across its input should be rather higher than 5.7 k Ω ; the combined value might be 10 k Ω , for example. This would mean using larger denominations for resistors R_1 and R_2 . This we may do, provided there is no objection to I_{E2} rising above 4 mA. Alternatively, one might try a different transformer ratio, such that Q_1 changes its position. We shall not go beyond this point. Our only reason for mentioning the complications is to show that it is often only after a certain amount of repeated recalculation that an acceptable compromise is arrived at and that all the requirements, which are often conflicting, are to some extent reconciled.

Use of a temperature-dependent resistor

Sometimes the element employed for R_1 is a resistor whose value decreases as its temperature rises — a thermistor¹²⁾ for example, or a germanium diode. The result of so doing is that E_B and R_B fall off with rising temperatures and the stabilization of I_E may be improved in consequence. Given R_2 one can calculate, for each value of t_j , what values R_1 must have in order that the transistor current shall keep to the prescribed value. In order to do this, the corresponding values of V_{EB} and I_B are

determined from the characteristics for the junction temperatures t_j in question. Formulae (3), (6) and (7) supply three equations involving the three unknowns E_B , R_B and R_1 . In this manner R_1 is determined as a function of t_j . Since I_E is now constant, V_C and P_C are constant too; therefore the difference between the transistor temperature and that of its surroundings is constant and, if K is given, this difference is known. If we assume that the temperature of resistor R_1 is always the same as the ambient temperature, we can say how R_1 should vary as function of the ambient temperature in order that I_E shall remain constant. A resistor or combination of resistors can now be sought which, exhibiting the desired type of dependence on temperature with reasonable precision¹³⁾, can satisfactorily be employed for R_1 . Once such a resistor or combination of resistors has been found, there will no longer be any need for the design to take account of the most unfavourable combination of temperature changes and at the same time to allow for shifts of characteristics due to spread. This gives one a wider margin for dealing with other requirements that the circuit has to satisfy.

Appendix: Physical background to temperature effects in transistors

The main reason for the profound effect of temperature on the behaviour of transistors is that, in semiconductors, the equilibrium concentrations of the minority charge-carriers are rapidly changing functions of the temperature. If there is equilibrium between the processes of generation and recombination¹⁴⁾, the product of the hole concentrations (p) and electron concentrations (n) in a semiconductor is given by

$$p_0 n_0 = K(T). \quad \dots \dots \dots (9)$$

The suffix 0 after p and n denotes that *equilibrium* concentrations are involved. $K(T)$ is an equilibrium constant such as is always involved in reversible reactions in physical chemistry. Here the reversible reaction is constituted by the generation of electrons and holes and their recombination. $K(T)$ is roughly proportional to $\exp(-V_g q/kT)$, where V_g represents the gap in eV between the valence and the conduction bands. Accordingly, qV_g is the energy required to liberate an electron, a hole being generated at the same time. Now, in an N region in a semiconductor, the electron concentration is virtually equal to the excess of donors over acceptors. The n_0 of such a region is therefore independent of temperature. In accordance with (9), therefore, p_0 must increase in proportion with $K(T)$, and hence in proportion with $\exp(-V_g q/kT)$. Conversely, in a P region it is n_0 that is proportional to $\exp(-V_g q/kT)$.

In order to make clear the influence that the equilibrium concentrations have on the properties of a transistor, we shall make recourse to the simple theory of the transistor dealt with

¹¹⁾ See for example I, page 20, left column.

¹²⁾ E. J. W. Verwey, P. W. Haaijman and F. C. Romeijn, Semiconductors with large negative temperature coefficient of resistance, Philips tech. Rev. 9, 239-248, 1947/48.

¹³⁾ A practical example may be found in C. Huber and J. Rodrigues de Miranda, A transistor pre-amplifier for the magnetodynamic pick-up, Philips tech. Rev. 18, 238-242, 1956/57.

¹⁴⁾ See for example J. C. van Vessel, The theory and construction of germanium diodes, Philips tech. Rev. 16, 213-224, 1954/55, especially pp. 214 and 215.

in an earlier article in this periodical¹⁵). Fig. 19 shows how the concentrations of minority charge-carriers are distributed inside a *P-N-P* transistor. In the junction planes 3 and 4 these minority concentrations are kept at zero by a potential difference of a few volts, applied in the reverse direction across the *P-N* junction between collector and base. Across the *P-N* junction between emitter and base a small voltage $V_{EB'}$ (approx. 0.1 V) is applied in the forward direction, and this keeps the minority concentrations in cross-sections 1 and 2 at values of $n_E(1)$ and $p_B(2)$, which are given by

$$n_E(1) = n_{E0} \exp(qV_{EB'}/kT), \quad \dots \quad (10a)$$

$$p_B(2) = p_{B0} \exp(qV_{EB'}/kT). \quad \dots \quad (10b)$$

n_{E0} and p_{B0} are the equilibrium concentrations of the minority charge-carriers in emitter and base respectively. Here the potential difference between emitter and base is allotted the symbol $V_{EB'}$. The prime sign indicates that the symbol refers to the voltage between the junction planes 1 and 2; this voltage is not the same as the voltage V_{EB} between the emitter and base electrodes because the base current, in passing through the transistor crystal, has to overcome what is termed the internal base resistance $R_{BB'}$ ¹⁶).

The gradient of the hole concentration in the base, and hence also the diffusion current of these holes, is proportional to $p_B(2)$. According to the simple theory as put forward in the article already referred to¹⁵, this diffusion current represents almost the whole of the transistor current I_E ¹⁷. Now V_g is approximately 0.75 V for germanium, and hence p_{B0} is proportional to $\exp(-0.75 q/kT)$. This in combination with (10b) gives us the following expression for the transistor current

$$I_E = C \exp[(-0.75 + V_{EB'})q/kT]. \quad \dots \quad (11)$$

C is a constant¹⁸) that evidently represents the current that

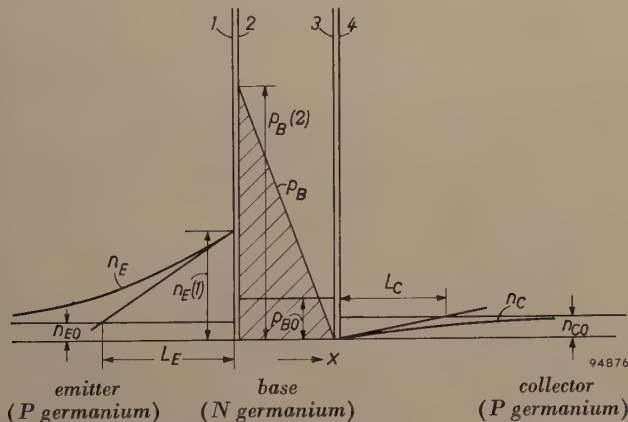


Fig. 19. Concentrations of minority charge-carriers as a function of position inside a *P-N-P* transistor; n_E are electrons in the emitter, n_C are electrons in the collector and p_B are holes in the base. The equilibrium concentrations of the minority charge-carriers in emitter, base and collector are n_{E0} , p_{B0} and n_{C0} respectively. The concentration gradients of the electrons in cross-sections 1 and 4 are $(n_E(1) - n_{E0})/L_E$ and n_{C0}/L_C respectively, L_E and L_C being the diffusion lengths in emitter and collector.

¹⁵) F. H. Stieltjes and L. J. Tummers, Simple theory of the junction transistor, Philips tech. Rev. 17, 233-246, 1955/56.

¹⁶) See II, page 103.

¹⁷) In actual fact, as the transistor current increases, the portion for which the electric field is responsible starts to become significant; see for example F. H. Stieltjes and L. J. Tummers, Behaviour of the transistor at high current densities, Philips tech. Rev. 18, 61-68, 1956/57, especially pp. 64 and 65.

¹⁸) C also exhibits some dependence on temperature, one reason being that the diffusion constant of the holes is temperature-dependent to some extent. However, so long as $V_{EB'}$ and T have practical values, this dependence is drowned by the effect T exercises on I_E via the exponential function.

would flow at a $V_{EB'}$ of 0.75 V if the simple theory continued to hold good with such large differences of potential across the junction¹⁹). Equation (11) shows that the I_E - $V_{EB'}$ characteristics having T as the running parameter are a family of exponential curves, all of which pass through point $I_E = C$, $V_{EB'} = 0.75$ V. If I_E is plotted logarithmically these characteristics become straight lines passing through the said point (fig. 20).

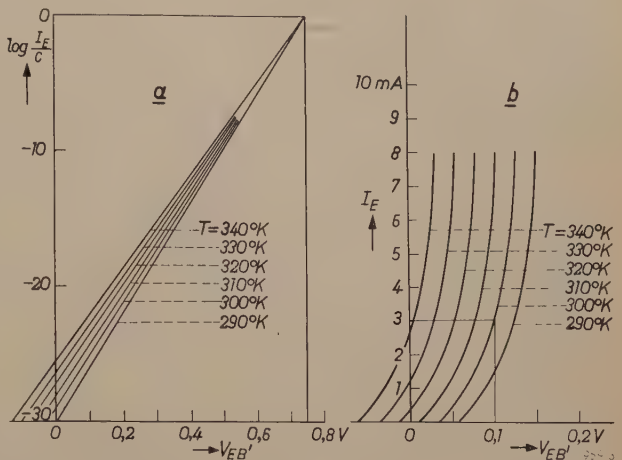


Fig. 20. Theoretical I_E - $V_{EB'}$ characteristics for a germanium transistor, with temperature as parameter. They are exponential in form (b) and they all meet at a point at which $V_{EB'} = 0.75$ V. This may be seen in graph (a), in which I_E is plotted logarithmically.

Equation (11) can be rewritten in the form:

$$V_{EB'} = -\left(\frac{k}{q} \ln \frac{C}{I_E}\right) T + 0.75. \quad \dots \quad (12)$$

It will be seen from this that if the current I_E is maintained at a constant value (say I_{E1}) when T changes, $V_{EB'}$ will decrease by a constant amount of $(k/q) \ln (C/I_{E1})$ per degree rise of temperature. If therefore a family of I_E - $V_{EB'}$ characteristics is drawn for a set of T values with the same interval (say 10 °K) between them, the curves will intersect a horizontal line $I_E = I_{E1}$ at equal intervals of $10 (k/q) \ln (C/I_{E1})$. The value of $(k/q) \ln (C/I_{E1})$ for a given I_{E1} cannot be calculated directly, since C is unknown. However, if I_{E1} is adjusted to a value such that $V_{EB'}$ is approximately 0.1 V at room temperature ($T \approx 300$ °K), then it follows from (12) that $(k/q) \ln (C/I_{E1}) = (0.75 - 0.1)/300 = 2.2 \times 10^{-3}$ V/°C. Thus we may conclude that the I_E - $V_{EB'}$ characteristic moves about 2.2 mV to the left for every °C rise in temperature. This is the first of the two rules given on page 128.

A complication arises on account of the internal base resistance $R_{BB'}$, which we have already had occasion to refer to in this appendix. The effect of $R_{BB'}$ is to make the I_E - V_{EB} characteristics rise less steeply than pure exponential functions; nevertheless their displacement per °C temperature rise remains the same. If the I_E - V_{EB} characteristic is only available for one temperature — a case that often arises in practice — good approximations to the true characteristics for other temperatures can be obtained by assuming that the entire curve moves to the left at the rate of about 2.2 mV/°C. With heavier operating-point currents, such as those in high-power transistors under Class A operation, the curve may move a good deal less, e.g. 2 mV/°C at $I_E \approx 0.5$ A and only 1 mV/°C at $I_E \approx 1$ A. Fig. 20a shows that this effect might have been anticipated on theoretical grounds. However, there are other effects, not discussed here, which also play a part.

¹⁹) The failure of the theory in these circumstances is discussed in the article cited in footnote 17).

In order to sketch in the background to the behaviour of the I_E - I_B characteristic, we may draw up a balance sheet for the electrons in the base region; the number entering it must be equal to the number leaving it plus the excess electrons undergoing recombination (i.e. the number of free electrons that are lost each second owing to recombination with holes in the base, minus the number gained in consequence of generation).

We shall make use of fig. 19 in drawing up our balance sheet. Electrons arrive in the base from the collector, the number thus gained each second being proportional to the concentration gradient in cross-section 4. In its turn, this gradient is proportional to n_{C0} , and hence to $\exp(-V_g q/kT)$. This, then, is the quantity to which the rate of arrival from the collector is proportional. The base current I_B is responsible for electron removal at the rate of I_B/q electrons per sec. Electrons are also transferred to the emitter, the number removed each second being proportional to the concentration gradient in cross-section 1. In its turn, this gradient is proportional to $n_E(1) - n_{E0}$ and hence — in accordance with (10a) — to $n_{E0}[\exp(qV_{EB'}/kT) - 1]$. It is generally permissible to neglect the term -1 in the square brackets. Since n_{E0} is proportional to $\exp(-V_g q/kT)$, we see that the rate at which electrons are transferred to the emitter is proportional to $\exp[(-V_g + V_{EB'})q/kT]$. The final item in our account is the loss due to excess recombination over generation. This is proportional to the area of the hatched triangle of fig. 19²⁰⁾ and hence to $p_B(2)$. But $p_B(2)$, like $n_E(1)$, is proportional to $\exp[(-V_g + V_{EB'})q/kT]$, and hence the loss due to excess recombination is also proportional to this expression. We now have the following for the rates at which electrons are gained and lost by the base:

Gain from collector $\propto \exp(-V_g q/kT)$ electrons/sec,
 Loss via I_B $= I_B/q$ " " "
 Loss to emitter $\propto \exp[(-V_g + V_{EB'})q/kT]$ " " "
 Recombination excess $\propto \exp[(-V_g + V_{EB'})q/kT]$ " " "
 This balance sheet of electron gain and loss gives us the following equation, in which the a 's are constants of proportionality:

$$a_1 \exp(-V_g q/kT) - (1/q)I_B - (a_2 + a_3) \exp[(-V_g + V_{EB'})q/kT] = 0. \quad (13)$$

By eliminating $\exp(-V_g + V_{EB'})q/kT$ from (11) and (13) and solving for I_E , we find that

$$I_E = \frac{-C}{q(a_2 + a_3)} I_B + \frac{Ca_1}{a_2 + a_3} \exp(-V_g q/kT). \quad (14)$$

This expression indicates that the I_E - I_B characteristic is a straight line that intersects the I_E axis at a value of I_E that is proportional to $\exp(-V_g q/kT)$. By differentiating with respect to T , we find that the intercept (which represents $-I_{CE0}$, cf. formula (2)) lengthens by $100 V_g q/kT^2$ per cent per degree of temperature. Since $V_g = 0.75$ V, $q = 1.6 \times 10^{-19}$ coulomb and $k = 1.38 \times 10^{-23}$ J per degree, the increase, in the region of room temperature ($T \approx 300$ °K), is about 10% per °C, i.e. the intercept doubles with every 7 °C. This is the second of the rules given on page 128.

Finally, let us compare I_E as function of T in the two cases of base current supply and voltage supply between base and emitter (see page 126 and fig. 10). By differentiating (14) and (11) with respect to T , we obtain the slopes of the curves for $I_B = \text{constant}$ and $V_{EB'} = \text{constant}$, respectively:

$$\left(\frac{\partial I_E}{\partial T}\right)_{I_B} = \frac{0.75 q}{kT^2} (-I_{CE0}),$$

$$\left(\frac{\partial I_E}{\partial T}\right)_{V_{EB'}} = \frac{(0.75 - V_{EB'})}{kT^2} I_E.$$

If we start from a point in the I_E - T plane at which $I_B = 0$, I_E is equal to $-I_{CE0}$ and, since $V_{EB'} \ll 0.75$, the slopes of

the curves for $I_B = \text{constant}$ and for $V_{EB'} = \text{constant}$ are about the same. Hence they more or less coincide. If however we start from a point at which $I_E \gg -I_{CE0}$, then

$$\left(\frac{\partial I_E}{\partial T}\right)_{V_{EB'}} \gg \left(\frac{\partial I_E}{\partial T}\right)_{I_B},$$

i.e. the curve for $V_{EB'} = \text{constant}$ rises much more steeply. In these circumstances, therefore, a base current supply is better than a voltage supply between base and emitter.

Instead of germanium, silicon is sometimes used for making transistors. V_g for silicon is higher than for germanium, being about 1.1 V instead of 0.75 V. In silicon, therefore, the effect of temperature on the equilibrium concentrations of the minority charge-carriers is even greater than it is in germanium. The I_E - V_{EB} characteristics of silicon transistors do in fact shift further per °C than those of germanium transistors. A calculation shows that, if $V_{EB'} = 0.1$ V, the displacement of the silicon transistor characteristic at room temperature will be $(1.1 - 0.1)/300 = 0.0033$ V/°C = 3.3 mV/°C. The percentage rise in $-I_{CE0}$ is also higher, being about 14.5% as against 10% per °C. Nevertheless, displacement of the I_E - I_B characteristics is more troublesome with germanium than with silicon transistors, because the value of $-I_{CE0}$ at room temperature is much smaller for the latter than for the former.

Summary. The biasing arrangements of a transistor, i.e. the network which determines the D.C. operating point, have a considerable bearing on the way I_E , the transistor current (which may be taken to mean the emitter or the collector current, as preferred), reacts to changes in temperature or to replacement of the transistor. The transistor current is sensitive to such changes on account of (1) displacement of the I_E - V_{EB} characteristic (V_{EB} being the potential difference between emitter and base) and (2) displacement of the I_E - I_B characteristic (I_B being the base current). These characteristics undergo displacement both when the temperature changes and when the existing transistor is replaced by another. With tubes, there is only one characteristic to be dealt with, namely the I_a - V_g characteristic, and since this is not affected by temperature changes, the situation is a good deal less complicated than that arising with transistors. Points of similarity and difference between transistors and tubes are noted throughout the article.

In a circuit biased by a voltage source between base and emitter, I_E is insensitive to displacement of the I_E - I_B characteristic but, on the other hand, highly sensitive to displacement of the I_E - V_{EB} characteristic. This is explained with the help of graphs. One can lessen the susceptibility of I_E to these changes by inserting a resistor R_E in the emitter lead or, alternatively, by inserting a resistor R_B in the base lead. However, the latter method renders I_E sensitive to displacement of the I_E - I_B characteristic. Raising the value of R_B to infinity (which results in base current biasing) renders I_E insensitive to displacement of the I_E - V_{EB} characteristic, but very sensitive to displacement of the I_E - I_B characteristic. Normally a combination of resistors R_E and R_B is employed.

In consequence of their sensitivity to temperature changes, the thermal equilibrium of transistors with their surroundings is of importance. The question of thermal equilibrium is investigated graphically. It is explained how a slight rise in ambient temperature can result in a large increase in transistor temperature (temperature "run-away"). A graphical method for analysing the behaviour of a given transistor in a given circuit is then discussed. The converse problem is also dealt with — the dimensioning (with the aid of graphs) of a circuit for a given transistor so as to satisfy certain requirements regarding thermal stability. The procedure is illustrated by the working out of an example; the circuit in question has in addition to allow for spread in the properties of different individuals of the same transistor type. Finally, the authors indicate how the sensitivity of I_E to temperature changes can be very largely compensated by including a resistor with a negative temperature coefficient in the circuit.

The physical background of the effect of temperature on transistor behaviour is treated in an appendix.

²⁰⁾ See the article cited in footnote ¹⁵⁾, page 240.

RADIO INTERFERENCE FROM FLUORESCENT LAMPS

by H. J. J. van BOORT, M. KLERK and A. A. KRUIHOF.

621.396.82:621.327.534.15

Radio reception, particularly of weak stations in the medium-wave band, is sometimes subject to interference from high-frequency oscillations generated in fluorescent lamps and entering the receiver via the mains. Quantitative information on this irregular phenomenon is obtained with a standard test arrangement, the results being analysed statistically. The authors describe an investigation on this basis into the strength of the interference penetrating into the mains with various types of lamps and ballasts. They indicate what they believe to be the cause of the interference and discuss some methods of reducing its penetration into the mains.

Capricious occurrence of the interference

It has been found that some fluorescent lighting installations interfere with radio reception. The interference is strikingly unpredictable in its occurrence, it being almost impossible to say beforehand whether it will arise with a given installation or not ¹⁾. As a rule the radio interference caused by the lamps constitutes a nuisance only at frequencies below about 1500 kc/s. It is hardly ever troublesome on the short-wave bands and quite inappreciable in frequency-modulation and television reception.

In this article we shall be concerned only with the long and medium waves, which are the most important regions affected by this form of interference. We shall investigate the reason for the irregular nature of the phenomenon and show how this irregularity can be largely eliminated for the purpose of measurements. Some quantitative and qualitative data on the origin of the interference will be given and methods indicated for reducing it.

The two causes of the irregularity

The irregular nature of the radio interference generated by fluorescent lamps has two entirely distinct causes. Firstly the oscillations produced in the lamp can fluctuate considerably in strength, and secondly the oscillations can be transmitted to the receiver in different ways, resulting in substantial variations in the coupling between the lamp and the receiver. We shall first consider the coupling between lamp and receiver and then describe a standardized method of measurement which, by eliminating the variation in coupling, gives a direct impression of the variations in the strength of the oscillations generated in the lamp.

Transmission of oscillations to the receiver

An article dealing at some length with the ways in which interference can enter a radio receiver was published in this journal as long ago as 1938 ²⁾. A brief recapitulation is given below of the contents of that article insofar as they are relevant to the present considerations.

The high-frequency oscillations produced in a fluorescent lamp can be transmitted along two different paths to a radio receiver: *a)* directly from the lamp to the aerial, and *b)* via the electric mains. The interference reaching the receiver by the direct path is noticeable only when the distance between the effective part of the aerial and the lamp is less than 2 to 3 metres. Since it should normally be possible when installing the receiver to make this distance larger (e.g. by using a screened lead-in cable) the direct lamp-to-aerial transmission of oscillations is left out of account in the standardized method of measurement. The transmission of interference via the mains involves capacitances which are generally disregarded in the more usual considerations of an electric-light mains, but which are by no means negligible where high-frequency oscillations are concerned. To illustrate the manner in which such oscillations are propagated from the lamp to the receiver via the electric-light mains, *fig. 1* shows schematically the principal paths through which high-frequency currents can flow ³⁾. In this diagram *p* and *q* represent the two mains conductors, and C_N , C_{N1} and C_{N2} the capacitances between these conductors mutually and with respect to earth. The fluorescent lamp *L* has a capacitance C_L with respect to earth. The capacitance of the coil of the ballast choke *S* is represented by C_S . The usual

¹⁾ A not inconsiderable proportion of the interference experienced is due to unreliable contacts in the electrical installation. Such interference is easy to remedy and is not dealt with in this article.

²⁾ L. Blok, Radio interference, Philips tech. Rev. **3**, 235-240, 1938.

³⁾ See also H. J. J. van Boort and M. Klerk, High-frequency oscillations in low-pressure mercury arcs, Appl. sci. Res. **B5**, 69-74, 1955.

suppressor capacitor C_0 is shunted across the glow-discharge starter G .

The part of the figure bounded by a broken line represents a small part of the receiver circuit. Of the

In the following we shall consider the symmetrical and asymmetrical components of the interfering voltage separately.

Fig. 3 shows schematically the capacitances from

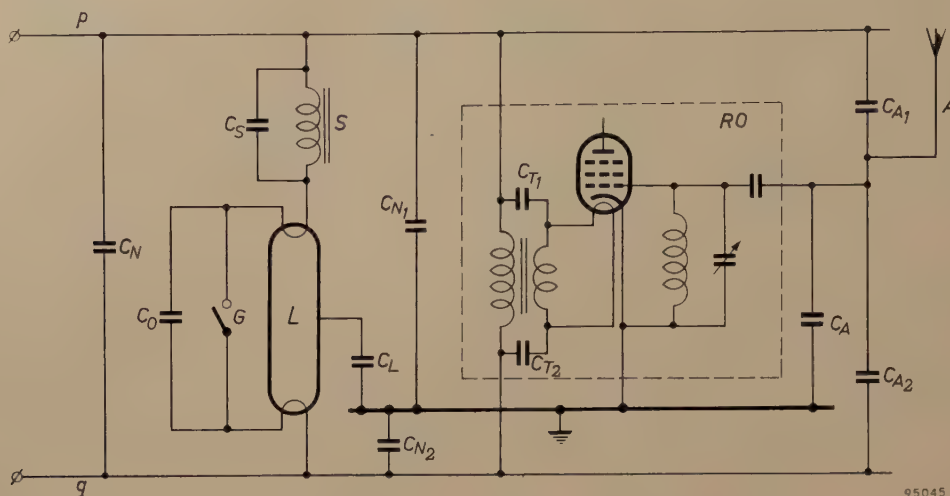


Fig. 1. Diagram of the principal circuit paths along which high-frequency electrical oscillations from a fluorescent lamp can enter a radio set via the mains conductors p and q . L lamp, G starter, C_0 suppressor capacitor, C_L capacitance between lamp and earth, S ballast choke, C_S self-capacitance of choke, C_N capacitance between mains conductors, C_{N1} and C_{N2} capacitances between mains conductors and earth, A aerial, C_A capacitance between aerial and earth, C_{A1} and C_{A2} capacitances between mains conductors and aerial, RO input stage of radio receiver, C_{T1} and C_{T2} capacitances via which the filament-current winding of the power transformer is coupled to the two mains conductors.

power transformer only the primary and the filament-current windings are shown. The latter is coupled to the mains conductors via the capacitances C_{T1} and C_{T2} between the windings. The aerial A has capacitances C_A , C_{A1} and C_{A2} with respect to earth and to the mains conductors.

It can be seen from this diagram that high-frequency oscillations can penetrate from the mains into the receiver in two ways, namely via the aerial and via the capacitances of the power transformer. In most practical cases the interference entering through the aerial is predominant.

When two voltages V_p and V_q of a given frequency are present on the mains conductors p and q , and these voltages differ both in amplitude and phase, each voltage can be resolved into two components, viz.

- 1) a component $+\frac{1}{2}V_1$ (for V_p) or $-\frac{1}{2}V_1$ (for V_q) consisting of voltages of equal amplitudes and opposite phase between the conductors and earth; V_1 is called the *symmetrical* component;
- 2) a component V_2 , consisting of voltages of equal amplitudes and the same phase between the conductors and earth, called the *asymmetrical* component.

This is illustrated by vector diagrams in fig. 2.

fig. 1 over which the high-frequency oscillations from the lamp can reach the aerial. The symmetrical and asymmetrical components of the high-frequency voltages on the lamp electrodes are denoted by V_1^* and V_2^* . They produce on the mains conductors p and q the symmetrical and asymmetrical voltages

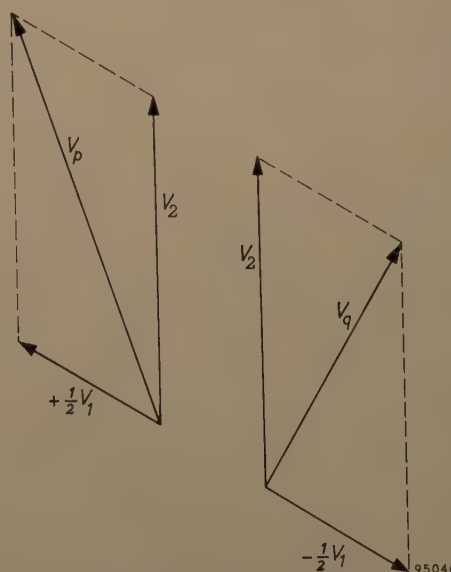


Fig. 2. The two alternating voltages, V_p and V_q , between the mains conductors and earth, resolved into a symmetrical component V_1 and an asymmetrical component V_2 .

V_1 and V_2 respectively. Since the circuit, generally speaking, is not symmetrical with respect to earth, V_1 is not exclusively caused by V_1^* , nor V_2 by V_2^* , but V_1 and V_2 are both influenced by V_1^* and V_2^* . In most cases, however, V_1^* is the main origin of V_1 , and V_2^* of V_2 .

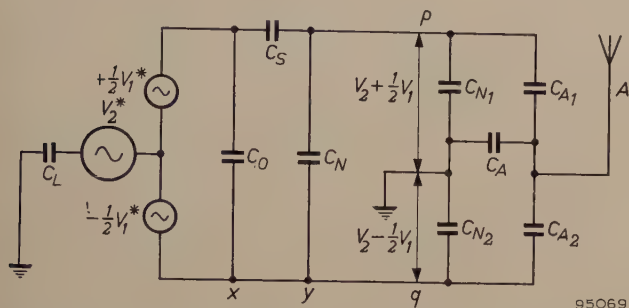


Fig. 3. Schematic representation of the circuit elements via which high-frequency oscillations from a fluorescent lamp can be propagated to the aerial. V_1^* and V_2^* represent the symmetrical and the asymmetrical components of the high-frequency voltages at the lamp electrodes. V_1 and V_2 are the corresponding high-frequency voltages on the mains conductors. The other letters have the same meaning as in fig. 1.

From fig. 3 we see that the symmetrical voltage component V_1 acts on a bridge circuit consisting of the capacitances C_{N1} , C_{N2} , C_{A1} and C_{A2} . If the bridge is balanced, i.e. if the condition $C_{A1}C_{N2} = C_{A2}C_{N1}$ is satisfied, the voltage V_1 will not cause any voltage to appear between the aerial and earth terminals. Although this condition is never exactly fulfilled, it is probable nevertheless that it has the effect of greatly reducing the influence of the symmetrical component of the interference voltage.

The current produced by V_2^* flows through C_L to earth and returns to the lamp along two principal paths, viz. via C_{N2} and via the series arrangement of C_{N1} and C_S . It is obvious that, as regards the asymmetrical component V_2 , it is not possible to achieve any marked reduction of the interfering voltage on the aerial terminals by balancing the bridge circuit on the right of fig. 3.

Concerning the penetration of interference into the receiver through the power transformer, it should be noted that several receiver elements are capacitively coupled to the two mains-conductors via the transformer. (In fig. 1 this is indicated for the filament winding by the capacitances C_{T1} and C_{T2} . Analogous capacitances are, of course, present in other windings of the transformer.) Consequently it is possible that here, too, the symmetrical component of the interfering voltage will not, or only very weakly, enter the receiver; for the asymmetrical component, there is again little chance of balancing out the interference voltages.

In practice all the capacitances in figs. 1 and 3

can have very divergent values, which means that the strength of the coupling between lamp and aerial is subject to considerable variation.

Measurement of the interference voltage

The Commission Internationale Spéciale des Perturbations Radiophoniques (C.I.S.P.R.) has standardized a method of measuring the symmetrical and asymmetrical voltage components introduced into the mains conductors by a source of interference. The circuit used is as shown in fig. 4. The interference source (in our case the fluorescent lamp) connected to terminals L is fed from the mains network N via chokes S_1 and S_2 . The interference current flows from L through the capacitors C_1 and C_2 and through the parallel branches of resistors. The values of the resistors are such that, for not too low frequencies, the impedance between the two mains conductors, and the impedance of both together with respect to earth, amount to 150 ohms.

The symmetrical component of the interference voltage is measured between terminals A_1 and A_2 , and the asymmetrical component between B_1 and B_2 . The measurements are performed with a standard test receiver of the super-heterodyne type. The input side, which is electrically symmetrical with respect to earth, contains an attenuator composed of resistors. The bandwidth of the intermediate-frequency section of the receiver is 10 kc/s. The output signal of the I.F. section is measured with a voltmeter, the response of which is made to simulate approximately the sensitivity of the human ear to crackling and hissing noises.⁴⁾

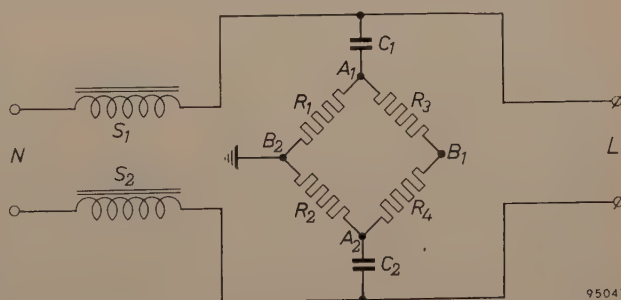


Fig. 4. Diagram of the circuit standardized by the C.I.S.P.R. for measuring the high-frequency oscillations transmitted by an interference source to the mains conductors. N mains, L interference source (fluorescent lamp with ballast), S_1 and S_2 isolating chokes, C_1 and C_2 coupling capacitors, capacitance $0.1 \mu F$, R_1 and R_2 resistors of 300Ω , R_3 and R_4 resistors of 100Ω . The symmetrical component is measured between points A_1 and A_2 , the asymmetrical between B_1 and B_2 (earth).

⁴⁾ This meter can be regarded as a peak voltmeter, but one which does not record voltage surges of very short duration or very small repetition frequency. This is achieved by a suitable choice of the capacitors and resistors and of the inertia of the meter.

With this set-up, then, it is possible to measure the two components of the interference voltage produced in the mains conductors by any interference source. Since in this arrangement the coupling between the interference source and the test receiver is fixed, the variation occurring in practice in the coupling between source and receiver is eliminated. The results of the measurements therefore provide information only on the voltage produced by an interference source, and do not directly reveal the nuisance it causes in a particular case. If measurements on diverse lamps disclose differences in the strength of the interference, these differences must stem from the lamps themselves.

Results of measurements on “TL” 40 W lamps

The most comprehensive interference measurements have been made on 40 W tubular (“TL”) fluorescent lamps. The ballast consisted of a choke which, for the purposes of the following, we shall refer to as type I. The suppressor capacitor used (C_0 in fig. 1) had a capacitance of 0.01 μ F. We shall first discuss the results of these tests; later on we shall indicate in how far the interference level differs from the values given here when other types of ballasts and lamps are used.

Altogether more than 800 “TL” 40 W lamps were investigated. They were obtained by daily setting aside, during a period of six months, a few lamps from normal production runs ⁵⁾. Before measurement the lamps were given two burn-in periods, each of 24 hours. Since the discharge sets in only at that end of each coiled filament electrode which is nearest to the mains, the lamps were reversed during the burning-in process so that the arc burnt equally long on both ends of the electrodes.

The measurements of the interference voltages were carried out at seven frequencies, viz. two in the long waveband, three in the medium wavebands and two in the transitional range in between. The frequencies are specified in Table I. No measurements were made in the short-wave bands, on which fluorescent lamps cause little or no interference.

The results of the measurements were similarly frequency-dependent in each of the three frequency ranges for all lamps; for this reason, only the mean values per frequency range are given in the Table. The results tabulated show first the mean values, S and A respectively of the symmetrical and asymmetrical components of the interference voltage,

both measured in decibels with respect to 1 μ V at the terminals of the test receiver. To give an impression of the variation about these mean values, the table also contains the standard deviation σ of both components ⁶⁾. Another important quantity is the level of the interference voltage which is not exceeded by “almost all” the lamps. If we take “almost all” to be 99% and if the results show a normal distribution around the mean, it can easily be shown that this quantity is equal to the larger of the two components, S , plus 2.3σ . This value is given in column M . Conversely, there is of course

Table I. Mean values of the symmetrical component S and of the asymmetrical component A of the interference voltage, measured with the C.I.S.P.R. test equipment. Both components are given in decibels above 1 μ V. σ is the standard deviation and $M = S + 2.3 \sigma$.

Frequency range	Measuring frequencies	S	A	σ	M
Long waveband	160 and 240 kc/s	52	38	3.1	59
Transitional region	350 and 550 kc/s	50	40	3.8	59
Medium waveband	750, 1000 and 1400 kc/s	42	33	4.9	53

a value of the symmetrical component that is exceeded by “almost all” the lamps, and that is equal to S minus 2.3σ . In a batch of about 100 lamps we can therefore expect a spread of 4.6σ in both components, and from the data in the Table we thus calculate the spread for each frequency range as 14, 18 and 23 dB, respectively. Furthermore, experience has shown that every lamp during its useful life can exhibit variations in its interference voltage of the same magnitude as found with a group of lamps all of the same age. The interference can both rise and fall during the life of a lamp, so that the average value in a large group remains much the same. Since there is a possibility that a lamp which initially causes little interference may later generate a great deal, one must always reckon with the maximum value M mentioned in Table I.

What level of interference is a nuisance?

It is evident that the nuisance experienced from a given level of interference voltages will be greater the weaker or more distant the station to which the receiver is tuned. In many countries the requirement, based on experience, is that both components of the interference voltage, measured with the C.I.S.P.R. test equipment, should be lower than 1 mV, i.e. less than 60 dB above 1 μ V.

⁵⁾ No systematic discrepancies were found between the production batches involved.

⁶⁾ Since the ratio between the two components in each of the three frequency ranges is almost constant, this standard deviation holds for both components.

In special cases it may be desirable to specify more stringent requirements, particularly for the reception of extremely weak stations, in laboratories where highly sensitive apparatus is in use, or where the impedances of the electric light mains have very unfavourable values, such as to prevent the partial compensation, described above, of the symmetrical component of the interfering voltage.

As we have seen, the value 60 dB is only very rarely exceeded in the case of "TL" 40 W lamps with a type I choke.

In order to discover whether the interference could be further reduced by making modifications in the lamp itself or in the circuit, it was desirable to ascertain at what point in the gas discharge the origin of the interference should be sought, and in what way the interference is connected with the mechanism of the discharge.

Origin of the interference in the lamp

For the purpose of locating the exact region in a lamp where the high-frequency oscillations arise, the lamp under investigation was fed via an isolating transformer, which made it possible to earth the circuit at any point. Also used for the investigation was a semi-circular metal strip, which we shall henceforth refer to as an "aerial probe"⁷⁾. This was fitted round the lamp and could be moved to and fro. The interference voltage picked up by this probe was passed to a C.I.S.P.R. test receiver, the I.F. signal of which was applied to the vertical

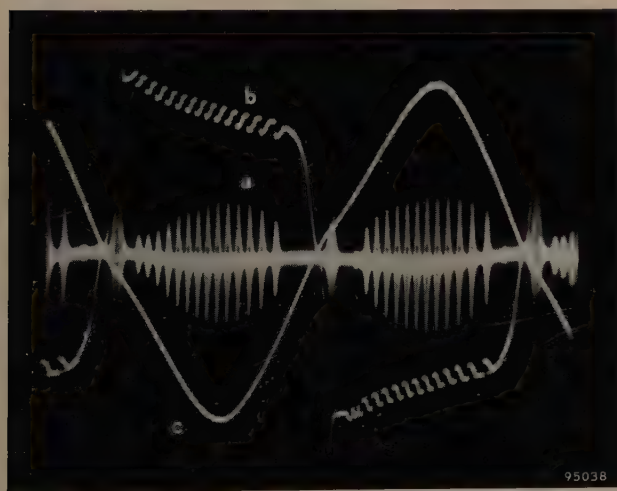


Fig. 5. Oscillograms of the interference voltage (a), the lamp voltage (b) and the current (c). The lamp circuit was not earthed. The aerial probe was fitted midway along the lamp, and the C.I.S.P.R. test receiver tuned to 300 kc/s.

⁷⁾ This investigation was begun round about 1942 by C. J. D. M. Verhagen, now professor of applied physics at Delft. See also the article quoted under ³⁾.

deflection plates of an oscilloscope. The horizontal deflection of the oscilloscope was synchronized with the mains voltage.

Fig. 5 shows the waveform (marked a) which appears on the oscilloscope screen when the lamp circuit is not earthed, the aerial probe placed midway along the lamp and the test receiver tuned to

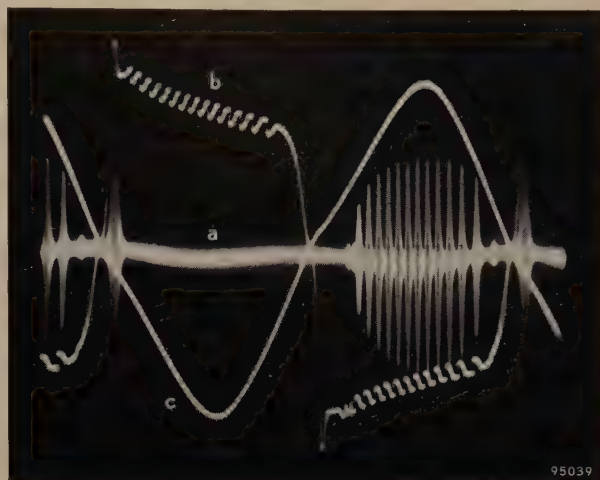


Fig. 6. As fig. 5, but now with one of the electrodes earthed and the probe placed near this electrode.

300 kc/s. The figure also shows the voltage (b) across the lamp, and the current (c). It can be seen that the high-frequency voltage a is very strong at the beginning of each half-cycle, just when the voltage across the lamp reaches the ignition potential. We shall call this part the *ignition interference*. Thereafter the interference voltage shows a number of peaks near the middle of each half-cycle. In many cases a strongish peak appears at the end of each half cycle (not visible in fig. 5); this we shall call the *extinction interference*. Usually the extinction interference is weaker than the ignition interference.

If we now earth one of the lamp electrodes and move the probe near that electrode, the waveform obtained on the oscilloscope screen is as shown in fig. 6. Obviously this waveform exhibits primarily that part of the oscillations which originates in the proximity of the earthed electrode, for the interference generated here gives rise to the full voltage over the small part of the discharge lying between the probe and the earthed electrode. Interference originating at the other electrode or at other positions in the lamp gives rise to only a small voltage over the same small part of the discharge.

From the absence in fig. 6, compared with fig. 5, of exactly half the oscillations, it can be concluded that the interference originates near the electrodes.

If we examine the polarity of the lamp we find that ignition and extinction oscillations are generated during the half-cycle in which the electrode concerned is the cathode. We shall therefore designate both as *cathode interference*. The oscillations in the middle of the half cycle arise when the electrode concerned is the anode, and these we shall refer to as *anode interference*⁸⁾.

The fact that both kinds of interference originate near the electrodes can be confirmed by other experiments. If the electrode where the aerial probe is positioned be heated by an additional current, the waveform of the cathode interference is found to be affected. A magnet brought close to the same electrode affects the waveform of the anode interference. The same experiments when carried out at the other end of the lamp produce no change in the oscillogram.

Tests with a movable probe mounted *inside* the lamp close to an electrode, demonstrated that the source of the cathode interference must be located at no more than a few millimetres away from that turn (or the two turns) of the coiled electrode where the discharge takes hold.

Further investigation showed that the cathode interference is chiefly made up of components with frequencies higher than about 350 kc/s, while the anode interference contains mainly components with frequencies lower than 250 kc/s. This is the reason that it was possible, for the statistical analysis of the results, to divide the entire frequency range into three parts according to the behaviour of the lamps. In the medium waveband the cathode interference predominates, in the long wavebands the anode interference, while in the transitional region now the one and now the other predominates.

Closer investigation of the cathode interference

The cathode interference, in view of its predominance in the important medium waveband, was subjected to a closer investigation, involving both the electrical and the optical phenomena occurring in a lamp.

In order to study the *electrical* phenomena the output signal of the I.F. section of a C.I.S.P.R. test receiver was again applied to the vertical deflection plates of an oscilloscope. For the horizontal deflection, use was now made of an extended time-base, enabling a small part of each cycle to be displayed over the whole breadth of the screen.

The *optical* phenomena can only be observed in lamps whose envelopes have not been provided with a fluorescent layer, for this layer scatters the light of the discharge to such an extent that the spatial distribution of the latter cannot be observed. Moreover, the inertia of the fluorescence phenomenon makes it impossible to observe rapid changes.

Two methods were adopted for studying the optical phenomena in a discharge lamp. In the first a photo-multiplier tube was mounted close to one of the electrodes and its output voltage applied to an oscilloscope. In the second the area around an electrode was observed through a narrow slit in a disc which was made to revolve synchronously with the mains frequency (fig. 7)⁹⁾. It was possible

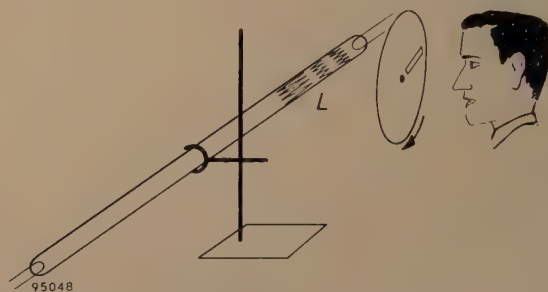


Fig. 7. Observation of the lamp *L* through a revolving disc *T*. Since the disc revolves in synchronism with the mains frequency, the observer views the discharge for a short instant of fixed phase during each cycle.

to vary the instant of the cycle at which the discharge was observed. In this way the observer was able to examine the discharge near an electrode for a very short time during each cycle. Owing to the inertia of the eye, however, no variations taking place *during* the instant of observation were perceptible. The photo-multiplier tube, on the other hand, has an almost instantaneous response to the total luminous flux emitted near the electrode under examination, and therefore provides a more detailed picture of the luminous intensity as a function of time.

Fig. 8 shows two oscillograms obtained with the aid of the aerial probe, the C.I.S.P.R. test receiver and the extended time-base. The middle waveform represents the ignition interference, the top and bottom waveforms the lamp voltage and current, respectively. Making allowance for a slight phase-shift in the equipment, one can conclude from these oscillograms that the ignition interference occurs at the moment when the voltage, after having passed

⁸⁾ Similar interference was found by H. L. Steele, *Illum. Engng.* **49**, 349, 1954, and by C. Matsuda, Y. Otami and R. Itadani, *Mem. Fac. Engng. Kyoto* **19**, 25-49, 1957.

⁹⁾ See also: J. W. Culp, *Noise in gaseous discharge lamps*, *Illum. Engng.* **47**, 37-46, 1952.

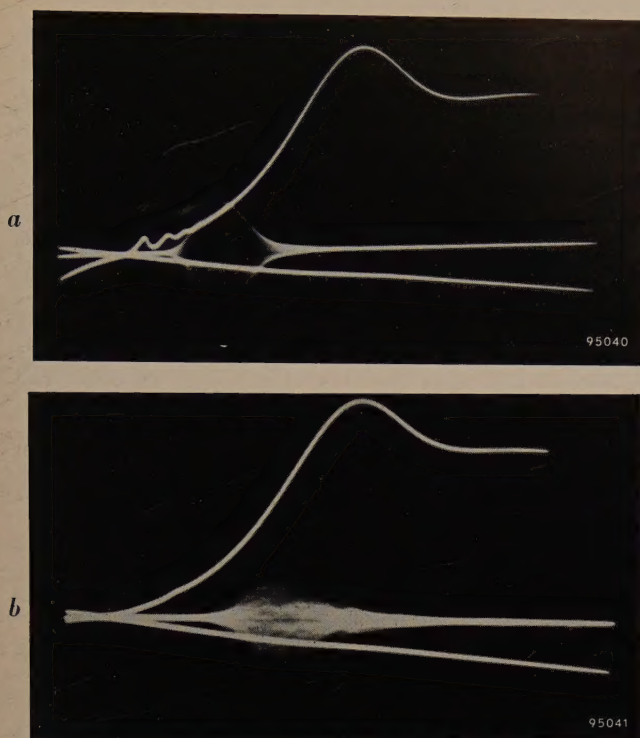


Fig. 8. Oscillograms of the ignition interference, recorded with the aid of an extended time-base. Upper waveform: lamp voltage. Middle waveform: interference voltage picked up in the test receiver via the aerial probe. A small phase shift occurs in the apparatus, so that the middle curve should be thought of as shifted about 4 mm to the left. Lower waveform: current. In *a* the interference consists of regular oscillations (stationary waveform envelope), in *b* of irregular oscillations.

through zero, again rises to about 14 V. It can also be seen that this interference can occur in two ways, sometimes as a regular oscillation with a stationary envelope, which also manifests itself in the voltage waveform, as in fig. 8*a*, but usually as an irregular oscillation (fig. 8*b*), in which case only a very faint kink is perceptible in the voltage waveform at the beginning of the oscillation. Analogous phenomena



Fig. 9. Oscillograms of the ignition interference with regular oscillations, recorded on expanded time base. Upper waveform: luminous intensity, recorded with photomultiplier tube. Lower waveform: lamp voltage.

are observed in the case of extinction interference.

The regular oscillations are found to have a frequency of some tens of kc/s. They are non-sinusoidal, so that harmonics are encountered at higher frequencies, but in the medium waveband these harmonics are quite weak. With the photo-multiplier tube the regular oscillations can also be very clearly observed in the light of the discharge. Fig. 9 shows an oscillogram (upper waveform) obtained in this way. As can clearly be seen, every maximum in the luminous intensity is really a double peak. In this case one observes through the revolving disc a luminosity that entirely surrounds the electrode (fig. 10*a*).

The strongest interference in the medium waveband is caused by lamps generating irregular cathode oscillations (fig. 8*b*). The frequency of these oscillations lies in the region of some hundreds of kc/s. Their amplitude is hardly affected by the external circuit. Through the revolving disc the luminosity is observed to occur at some distance from the electrode early on in the cycle (fig. 10*b*) and only later in the cycle surrounds the electrode entirely.

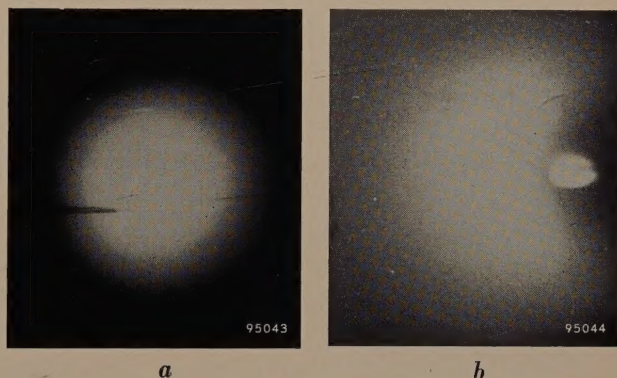


Fig. 10. *a*) Luminosity near the cathode during ignition interference with regular oscillations. The glowing cathode is in the middle of the luminous area (but not visible owing to over-exposure of the photo). *b*) The same during ignition interference with irregular oscillations. The glowing cathode is visible on the right of the discharge.

The irregular oscillations can also be generated in a lamp operating on direct voltage. To achieve this, the discharge current must be smaller than the (temperature-dependent) cathode saturation current. Plainly, in an A.C. discharge this state occurs at the beginning and end of each half cycle in which the electrode concerned is the cathode, for owing to the thermal inertia of the electrodes the saturation current at the beginning and end of each cathodic half-cycle is still fairly high, in spite of the fact that heating chiefly takes place about midway through each half cycle. Since the lamp current gradually rises and falls, the state favouring

the generation of irregular oscillations will occur twice per cycle for each electrode.

The phenomena described suggest that the irregular oscillations may be regarded as oscillations of positive ions in the potential minimum which occurs in front of the cathode while the discharge current is lower than the saturation current of the cathode.

Since a high saturation current is necessarily associated with the cathode condition needed to ensure that the lamp has an adequate life, it is not likely that the level of the oscillations generated in the lamp can be appreciably lowered without impairing the desired properties of the cathode. When it is required to reduce the radio interference caused by the lamps, it is therefore necessary to find means of preventing the oscillations from entering the electric mains.

Means of suppressing the interference

It should be noted first of all that the choke in the ballast and the capacitor C_0 in fig. 1 should both be regarded as means of reducing the interference. The self-capacitance C_S of the choke constitutes the largest impedance in the principal path for the symmetrical component (see fig. 3). This is confirmed by the fact that when a capacitor is shunted across the choke the symmetrical component shows a marked increase. With a capacitor of 1250 pF the respective increases in the long-wave, transitional and medium-wave bands were found to be 16 dB, 12 dB and 5 dB. By splitting the choke into two parts, and including one of these parts in each of the leads of the lamp it is possible to reduce the symmetrical component. The reduction obtained in a specific case was 12 dB. This component can be reduced by about 20 dB, if the choke be wound according to a special system which leaves both halves with a particularly low self-capacitance.

In the circuit path for the asymmetrical component the largest impedance is usually the capacitance C_L between the lamp and earth (see fig. 3). Increasing this capacitance by earthing the fixture on which the lamp is mounted, or by earthing a metal fixture around the lamp, is found to increase the asymmetrical component by 12 and 20 dB respectively. If there is a choke in each of the two leads to the lamp, the impedances represented by the self-capacitance of these chokes also affect the asymmetrical component. (In that case a capacitor C_S is inserted in the lead between points x and y in fig. 3, so that the impedance in this branch is increased.) In this way we were able to reduce

the asymmetrical component by 4 dB. With a low-capacitance system of winding the reduction amounted to 14 dB. The impedance between the lamp and the mains can be further increased by introducing low-capacitance high-frequency chokes in the two leads to the lamp. With chokes having a self-capacitance of only 15 pF it was possible to reduce the asymmetrical component by about 18 dB. Often, however, the leads from the choke to the lamp run parallel with the mains conductors, which means that the high-frequency chokes are bypassed by capacitances which can be much larger than their own self-capacitance. In such cases nothing is gained by using these chokes.

The suppressor capacitor C_0 reduces both components of the interference voltage, but just as with C_S , its effect is greater on the symmetrical component. If C_0 is removed, the symmetrical component increases by about 20 dB and the asymmetrical component by about 10 dB. Increasing C_0 from 10 000 to 20 000 pF causes a drop of 5 dB in the symmetrical and 2 dB in the asymmetrical component.

A simple method of further suppressing the interference is to use a so-called delta filter, consisting of three capacitors connected as shown in fig. 11.

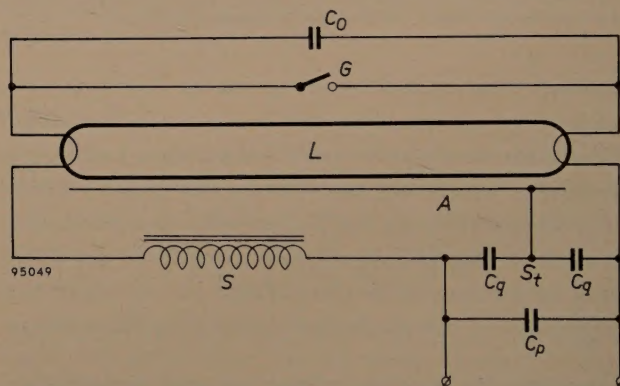


Fig. 11. Fluorescent lamp with series-connected choke and delta filter. L lamp, G glow-discharge starter, C_0 suppressor capacitor, S choke. The delta filter consists of three capacitors: $C_p = 0.2 \mu\text{F}$ and $C_q = 2500 \text{ pF}$. The star point St is connected to the fixture A . The fixture must not be earthed.

The capacitor C_p , of $0.2 \mu\text{F}$, is connected between the two mains conductors, hence in parallel with C_N (see fig. 3). This results in a drastic drop, about 35 dB, in the symmetrical component. The remaining magnitude of this component is roughly inversely proportional to C_p^{10} .

The mains conductors are connected via the two capacitors C_q to the fixture A . To eliminate the

¹⁰⁾ It should be noted that the shunt capacitor which can be used to improve the power factor $\cos \varphi$, also causes, like C_p , a marked drop in the symmetrical component.

possibility of receiving an electric shock from the fixture, the capacitance of each must not exceed 2500 pF. They influence almost exclusively the asymmetrical component, the more so the better the lamp is shielded by the fixture.

In Table II some figures are given of the reduction in the asymmetrical component brought about by the use of a delta filter. Column 1 gives the reduction in dB in the earlier-mentioned three frequency ranges when the lamp was mounted at a distance of 2 cm from a metal fixture with a breadth of 3.5 cm. The fixture was connected to the star point *St* of the filter (see fig. 11). Column 2 gives the reduction obtained after a metal cover had been fixed above the lamp.

Table II. Reduction, in dB, obtained in the asymmetrical interference component by the use of a delta filter. Column 1: "TL" lamp mounted on metal fixture. Column 2: after fixing a metal cover above the lamp, mounted on fixture.

Frequency range	1	2
Long waveband	10	22
Transitional region	6	15
Medium waveband	4	17

It appears from this table that a metal shield at one side of the lamp, combined with a delta filter, causes appreciable suppression of the interference. The same degree of suppression is achieved when, instead of a metal cover, metal louvres are fixed underneath the lamp, again connected to the starpoint *St*. An advantage of using a delta filter is that no earth lead is necessary. In many cases it is not possible to install a good earth lead, and in other cases the lead would have to be so long that it would itself act as a coupling element with the receiver aerial. If a good, short earth lead is available, it may then be beneficial to connect the starpoint of the delta filter to it, but *not* the fixture, for that would mean an increase in C_L (see fig. 3).

Other ballasts and other „TL” lamps

The investigation described in this article has been carried out over a period of some years. The choke referred to as type I, and which was used for the experiments hitherto discussed, was of a kind manufactured up to about 1951. Tests were subsequently carried out with a later version (type II, known as "Alfo" ballasts¹¹⁾, manufactured up to about 1956, in which the filler used between

core, coil and casing was crinkled aluminium foil. Finally, interference measurements were also made with the latest ballasts (type III), containing a filler on a polyester-resin base¹¹⁾.

Since the question obviously arose whether the interference might perhaps enter the mains conductors to a different extent when other types of ballast were used, tests were also made on "TL" 40 W lamps with capacitive ballasts, consisting of a choke and capacitor in series. These ballasts are employed in large installations to improve the power factor.

Although far fewer tests were carried out in this investigation than in the experiments discussed in the foregoing, they were enough to provide a general picture of the level of interference arising with the diverse lamps and ballasts.

The averages of the results of these tests for the three frequency regions are given in Table III, together with the corresponding figures for type I, taken from Table I, for comparison.

The spread in the results for a large number of lamps is independent of the ballast, so that the values of the standard deviation σ , given in Table I, also apply to the other ballasts.

We see from Table III that in the important medium-wave bands the larger component of the interference voltage is smaller for the choke ballasts types II and III and the capacitive ballast than for the choke of type I.

Table III. Average values *S* and *A* of the symmetrical and asymmetrical components, respectively, of the interference voltage using different ballasts, viz. chokes of types I, II and III and a capacitive ballast, with a "TL" 40 W lamp. Both components are given in dB above 1 μ V.

Frequency range	Type I		Type II		Type III		Cap.	
	<i>S</i>	<i>A</i>	<i>S</i>	<i>A</i>	<i>S</i>	<i>A</i>	<i>S</i>	<i>A</i>
Long waveband	52	38	52	38	47	43	49	35
Transitional region	50	40	44	37	40	41	45	38
Medium waveband	42	33	34	30	33	34	37	33

In many cases an inductive and a capacitive ballast are combined to form what is called a "duo-ballast". The interference from two "TL" 40 W lamps with a duo ballast is found to be no greater than that from one lamp.

If two "TL" 20 W lamps connected in series be substituted for one "TL" 40 W lamp, no change is found in the asymmetrical component, but the symmetrical component increases by about 2.5 dB; it increases by twice this amount when four 20 W lamps are substituted for the two 40 W lamps used with a duo-ballast. Longer lamps and lamps of higher current generally cause somewhat stronger

¹¹⁾ Th. Hehenkamp, Small ballasts for fluorescent lamps, Philips tech. Rev. 18, 279-281, 1956/57.

interference, whereas shorter lamps and lamps of lower current cause somewhat less; the differences, however, are not very large.

Table IV gives the average results of tests performed on two other types of lamp, viz. "TL" M lamps with the appropriate ballast¹²⁾ and "TL" S lamps in series with an incandescent lamp¹³⁾.

Table IV. Mean values of the symmetrical component S and the asymmetric component A of the interference voltage from "TL" M and "TL" S lamps. Both components are expressed in dB above 1 μ V.

Frequency range	"TL" M		"TL" S	
	S	A	S	A
Long waveband	42	44	80	53
Transitional region	40	42	78	55
Medium waveband	30	32	70	48

It can be seen that with the "TL" M lamps the asymmetrical component is slightly larger than the symmetrical. Both are well below the specified limit of 60 dB.

In the case of "TL" S lamps in series with an incandescent lamp, both components are much larger than with the lamps discussed in the foregoing. We see that the symmetrical component is considerably in excess of 60 dB. Two reasons can be indicated for this higher level of interference:

- a) Since there is no choke in the lamp circuit, the current in "TL" S lamps is subject to more abrupt variations, causing both components at all frequencies to increase by about 8 dB above those in a circuit with a choke ballast.
- b) The suppressor capacitor C_0 (see fig. 1) is not generally used with "TL" S lamps in series with an incandescent lamp. This results in an increase of

20 dB in the symmetrical and 7 dB in the asymmetrical component of the interference voltage.

In order to keep the interference from "TL" S lamps below the usual 60 dB limit it is essential to use a delta filter. This reduces the symmetrical component by 31 dB, resulting in a level somewhat lower than that of a "TL" 40 W lamp with choke and suppressor capacitor. The asymmetrical component is only slightly reduced by a delta filter when no shielding fixture is used; nevertheless this component is then just small enough to make the cases in which the 60 dB limit is exceeded a rare occurrence. If necessary the symmetrical component can be further reduced by increasing the capacitance C_p of the delta filter, and the asymmetrical component can be lowered by electrical screening by a metal fitting.

The interference from "TL" S lamps operated with a ballast containing a choke and suppressor capacitor is, of course, no greater than from a normal "TL" lamp with the same ballast (see Table III).

Summary. The radio interference from fluorescent lighting installations is a capricious phenomenon. The irregularity is due to the marked variations in the strength of the interference source and of the coupling between the lamp and the receiver. The interference enters the receiver principally via the electric-light mains and the aerial. A test arrangement standardized by the C.I.S.P.R. is used for measuring the symmetrical and asymmetrical components of the interfering voltage induced in the mains conductors by a fluorescent lamp with ballast. The usual upper limit specified for both components is 1 mV, i.e. 60 dB above 1 μ V. It is found that this limit is appreciably exceeded only with "TL" S lamps when series-connected with an incandescent lamp and with no suppressor capacitor (in the absence of any special measures of suppression).

The interference source is located in the vicinity of the lamp electrodes. The strongest interfering voltage is generated near an electrode during that part of each cycle in which it functions as cathode. The cause of the interference is considered to be oscillations of positive ions in the potential minimum which occurs in front of the cathode when the discharge current is smaller than the temperature-dependent saturation current of the cathode. Means of reducing the interference penetrating into the mains conductors are: splitting the choke into two parts, using low-capacitance chokes, electric shielding of the lamp and using a three-capacitor delta filter. With the latter the interference from "TL" S lamps in series with incandescent lamps can also be kept below the permissible limit.

¹²⁾ W. Elenbaas and Th. Hehenkamp, A new fluorescent lamp in a starterless circuit, Philips tech. Rev. 17, 288-293, 1955/56.

¹³⁾ W. Elenbaas and T. Holmes, An instant-starting fluorescent lamp in series with an incandescent lamp, Philips tech. Rev. 12, 129-135, 1950/51.